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Valve Microphony

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The first part of this article considers in general terms the production of microphony. The various methods of investigating microphony are discussed, and their advantages and disadvantages considered. Illustrations of microphony in various valves are given. The second part considers in more detail microphony in radio and television receivers. Calculations are made of the permissible levels of microphony in the various stages of receivers.

Part 1—Production of Microphony and Methods of Investigation

INTRODUCTION

Many components in electronic equipment do not have a completely rigid structure but consist of parts that can vibrate when the component is subjected to physical excitation. As the parts vibrate, the distance between them may vary and variations in the electrical properties of the component can occur. As a simple example, if the plates of a tuning capacitor vibrate, the distance between the plates may vary and, as a result, there will be a corresponding variation of the capacitance. If the capacitor forms part of the tuned circuit of an oscillator, the frequency of oscillation will vary correspondingly. In other words, the vibration of the capacitor plates with respect to one another gives rise to an interference signal which produces frequency modulation of the oscillator signal. The production of an interference signal as a result of mechanical vibration is known as 'microphony'.

This article considers microphony in valves. The vibration of the electrode structure of the valve can cause variations not only in the inter-electrode capacitances, but also in the anode current and mutual conductance. The vibration may be a result of shocks — for example, equipment in cars or aircraft, and equipment containing push buttons or stiff switches will be subjected to this type of excitation — or the result of continuous excitation from the motors of gramophones and tape recorders, and from the loudspeakers of radio and television receivers. The loudspeaker is often the most troublesome cause of microphony since it is placed close to the valves and can transmit vibrations to the valves both mechanically (through the cabinet and chassis) and acoustically (through the air). Besides causing sound interference, the excitation of the loudspeaker can also affect the valves in the picture channel or timebase circuits of a television receiver and produce interference on the picture. In television receivers, picture microphony is usually more troublesome than sound microphony.

In spite of the very narrow tolerances used in the manufacture of valves, it is impossible to avoid slight differences between valves of the same type. These differences do not have any significant effect on the electrical characteristics of the valve but may affect the microphony considerably. Consequently, certain methods of testing valves can be carried out only on a statistical basis and any modification to a valve to reduce its microphony can be checked only by testing a large number of valves.

FACTORS DETERMINING STRENGTH OF MICROPHONY

A valve subjected to acoustical or mechanical vibration undergoes a varying acceleration. Primarily, it is the magnitude and frequency of this acceleration that determines the strength of the microphony. As an example of the magnitude of the accelerations involved, a typical accelerometer trace is shown in Fig. 1. This trace, of acceleration plotted against excitation frequency, was obtained by replacing one of the valves in a radio receiver by an accelerometer and driving the loudspeaker with a constant 50mW signal of varying frequency. Generally, peak accelerations of between 0·1 to 0·2g (g is the acceleration caused by gravity) can be obtained with loudspeaker powers of 50mW. The magnitude of the acceleration increases with an increase in excitation power, the increase being approximately proportional to the square root of the power.

It has already been mentioned that the vibration may be transmitted from the source to the valve mechanically or acoustically. In the case of mechanical transmission with vibrations reaching the valve through the cabinet and chassis, the magnitude of the acceleration of the valve will depend on such factors as the rigidity of the valve in its holder and the rigidity of the valve holder in the chassis. For acoustic transmission, the main factor affecting the acceleration of the valve is the extent to which the vibrations are transmitted through the valve envelope. The factors affecting the magnitude of the acceleration of the electrodes are the acceleration of the valve, the physical properties of the electrodes, the stiffness and rigidity of the electrode structure, and the method of mounting inside the valve.

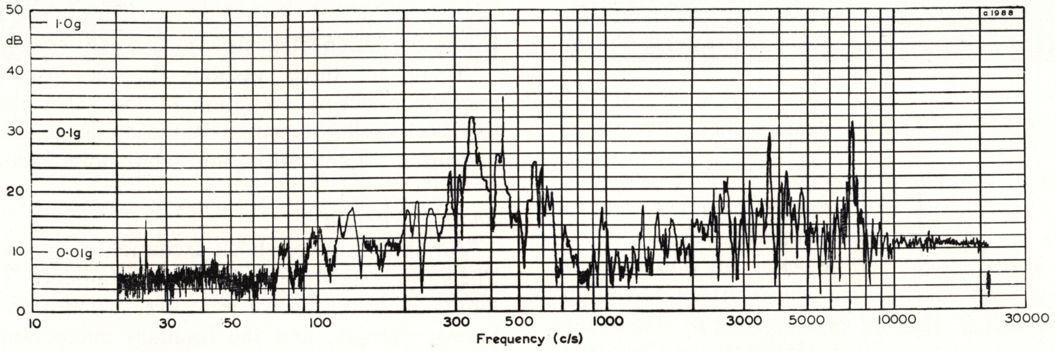

Fig. 1—Pen trace of acceleration plotted against frequency for a valve in a radio receiver subjected to a constant loudspeaker power of 50mW

Further points that have to be considered in connection with the strength of the microphony are the application, the sensitivity, and the operating conditions of the valve. These points are considered in more detail in Part 2 of this article.

INCONSISTENT NATURE OF MICROPHONY

Because of the many factors determining its strength, microphony is very inconsistent. The inevitable slight differences between valves of the same type already mentioned can cause differences in microphony between valves of the same type and valves in the same application. Again, a particular valve may suffer from microphony in one application but not in another, and the position of the valve on the chassis and the method of mounting may affect the microphony considerably. The acceleration on the valve is extremely irregular, as shown by the trace of Fig. 1, since the chassis, cabinet, and other structural members of the equipment have many different resonant frequencies for mechanical and acoustical vibrations, and the whole structure behaves as a combination of mutually coupled resonators. Consequently a small structural change in the equipment can completely alter the frequency spectrum of the acceleration of the valve. For these reasons, it is not possible to calculate accurately the accelerations to which the valve will be subjected, but these can be measured by means of the accelerometers shown in Fig. 2. These laboratory-made accelerometers enable the acceleration on the valve to be measured in three mutually perpendicular directions. The accelerometers are similar in weight and shape to a valve and can be inserted in a valve holder in the equipment under test.

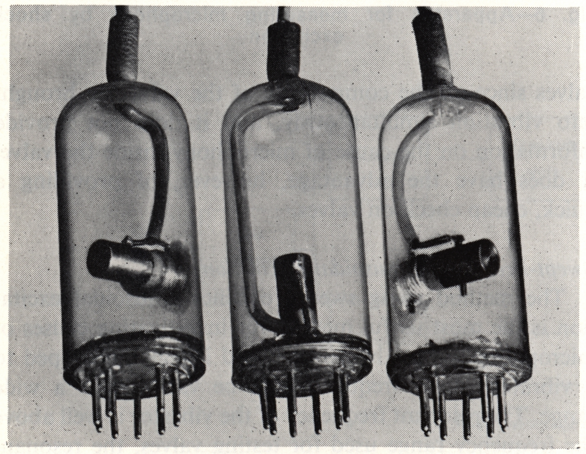

Fig. 2—Accelerometers for measuring the acceleration of a valve in three mutually perpendicular directions

METHODS OF INVESTIGATING MICROPHONY

The methods for investigating microphony can be classified into two groups. The first group uses indirect excitation of the valve such as loudspeaker excitation through the cabinet and chassis. The second group of methods uses direct excitation of the valve itself.

Indirect Excitation Methods

The microphony of an audio amplifier valve may be investigated using the method shown in Fig. 3. The valve is incorporated in an amplifier stage of variable gain and placed near the loudspeaker. The gain of the amplifier

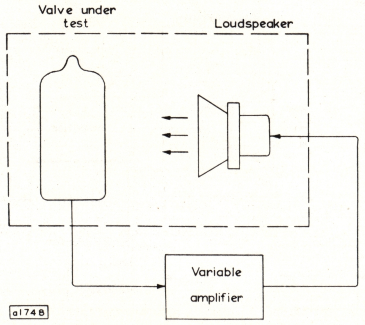

Fig. 3—Method of measuring a.f. microphony by indirect excitation of the valve

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Part 1 of this article is an abridgement of 'Microphony in Electron Tubes' by S. S. Dagpunar, E. G. Meerburg and A. Stecker, published in **Philips Technical Review**, Vol. 22, No. 3, January 1960.

is increased until acoustic feedback (or feedback howl) occurs. The gain of the amplifier is then gradually reduced until the howl is just audible. The gain control is left in this position, and the sensitivity of the valve for just-audible howl is determined by injecting a signal of known amplitude into the valve and noting the loudspeaker power. The sensitivity is generally defined as the voltage on the grid of the valve required to produce a loudspeaker power of 50mW. The sensitivity for just-audible feedback howl gives an indication of the maximum amplification of the valve that is permissible.

The performance of an r.f. valve can be tested in a similar way, as is shown in Fig. 4. An unmodulated r.f. signal is fed to the control grid of the valve and a detector connected to the output. The output of the detector is fed to the loudspeaker through a variable amplifier. The microphony of the valve causes modulation of the r.f. signal, and the detector delivers an audio-frequency signal which is amplified sufficiently to produce feedback howl. The sensitivity of the valve for just-audible feedback howl can then be determined as before. This sensitivity is defined as the modulation depth required to produce a loudspeaker power of 50mW.

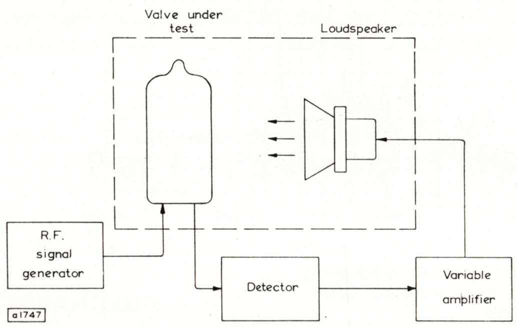


Fig. 4—Method of measuring r.f. microphony by indirect excitation

In a similar way, the microphonic tendency of the oscillator valve can be found by determining the sensitivity for just-audible feedback howl. In this case the sensitivity is defined as the frequency deviation required to produce a loudspeaker power of 50mW.

Another method of testing for microphony is shown in Fig. 5. The loudspeaker is excited at a swept frequency by an a.f. oscillator and amplifier, and the microphony at the desired frequency is measured by an amplifier and valve voltmeter.

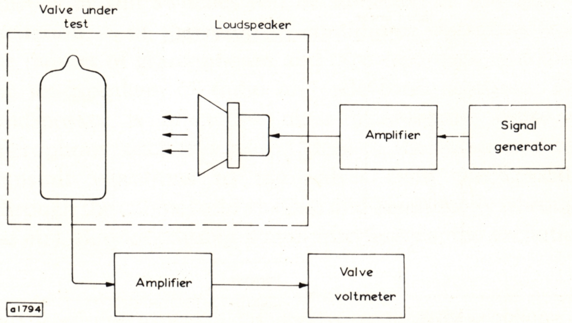


Fig. 5—Method of measuring microphony at a particular frequency by indirect excitation

However, as was shown in Fig. 1, even when the loudspeaker power is constant the acceleration on a valve is very irregular and, as a result, it is not easy to investigate the microphony characteristics of the valve itself since there is always the possibility that the cause of the microphony lies outside the valve. To obtain results that refer solely to the valve, methods that excite the valve directly and not through a loudspeaker and chassis must be used.

Direct Excitation Methods

There are three methods for the investigation of microphony using excitation of the valve alone. These use shock excitation, swept-frequency sinusoidal excitation at constant acceleration, and white-noise excitation.

Shock Excitation

An apparatus for shock excitation is shown in Fig. 6. The valve is incorporated in an a.f. amplifier. The pendulum is allowed to strike the valve, giving it an impact of known strength, and the resultant microphonic voltage measured. The method is of use only for comparing valves since all the components of the valve are brought into vibration simultaneously, and it does not provide information on the cause of microphony inside the valve. It does have the advantage, however, of providing a quick, cheap check on valves.

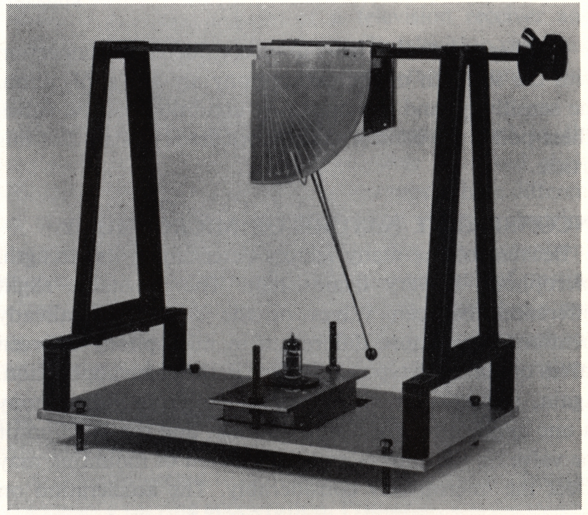


Fig. 6—Apparatus for measuring microphony by shock excitation

Swept-Frequency Sinusoidal Excitation

The method using swept-frequency sinusoidal excitation is the most useful method for investigating causes of microphony. The valve to be tested is rigidly clamped to a vibrator whose frequency can be varied over a wide range. The resonant frequency of the vibrator is well above the frequency range used for testing valves, the resonant frequency being about 30kc/s and the testing range 30c/s to 20kc/s. An accelerometer is mounted on the vibrator to monitor the acceleration on the valve. Photographs of valves mounted on the vibrator are shown in Fig. 7.

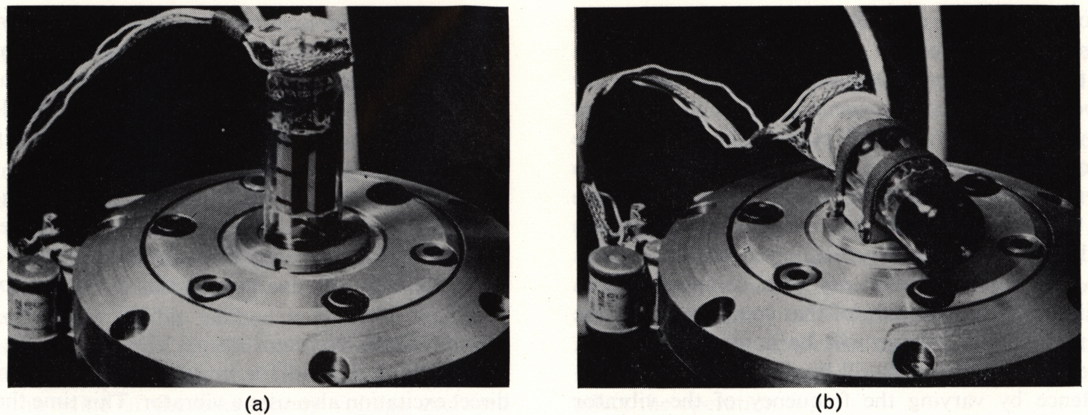


Fig. 7—Valve mounted on the vibrator:
(a) with vibration along longitudinal axis of valve
(b) with vibration perpendicular to longitudinal axis of valve

Interference voltages resulting from the vibration are measured as the frequency of the vibrator is slowly varied. Typical examples of pen recordings of microphonic voltage plotted against frequency are shown in Figs. 18 to 20. The peaks in the trace correspond to the resonant frequencies of various parts of the electrode structure of the valve. The part of the structure responsible for the microphony is best discovered by examining the valve with a microscope and stroboscope, and a block diagram of the equipment for this is shown in Fig. 8.

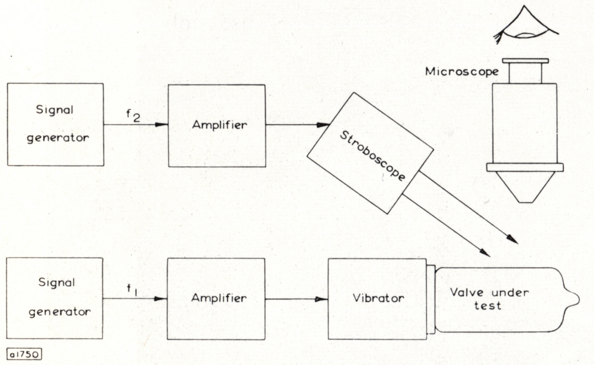


Fig. 8—Block diagram of equipment for examining vibrating valve

Two signal generators are used, one to operate the vibrator and the other to operate the stroboscope. The difference in frequency between the two generators is kept small—a few cycles per second—so that the vibrating parts of the electrode structure appear to be moving very slowly when viewed through the microscope. It is difficult in practice to couple the frequency controls of two signal generators to give a small frequency difference over a wide range and in Fig. 9 a block diagram is shown of the practical version of the equipment. In this case one signal generator is used, feeding the vibrator directly through an amplifier and the stroboscope through a frequency-shift network. This network introduces a constant frequency difference of a few cycles per second between the frequency of the signal generator and that of the signal fed to the pulse generator and stroboscope. The output voltage from the valve is taken to one pair of deflection plates of the oscilloscope and the other pair of plates is supplied with a voltage proportional to the current driving the vibrator.

As the frequency of the vibrator is varied, so the microphonic output voltage of the valve will vary. With any vibrating system, if the frequency of the driving force is lower than the resonant frequency of the system, the deflection will be in phase with the force. If the frequency

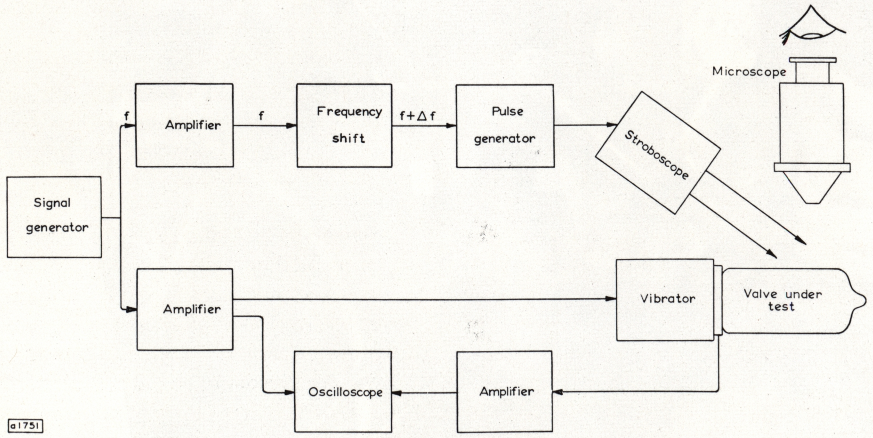


Fig. 9—Block diagram of practical version of equipment shown in Fig. 8

of the force is higher than the resonant frequency, the deflection and force are 180° out of phase. The transition between these two states takes place in a narrow frequency range about the resonant frequency, and at the resonant frequency itself there is a 90° phase shift between the deflection and the force. The changes of phase will also be shown on the oscilloscope as the phase difference between the current driving the vibrator and the voltage produced by the microphony, displayed as a Lissajous figure. At the resonant frequency of a component both the amplitude of the vibration of that component and the microphonic output voltage will be a maximum. Also, the sudden change of phase will be shown on the oscilloscope. Hence by varying the frequency of the vibrator until the microphonic voltage shows a maximum, and by observing the oscilloscope trace and the valve through the microscope, it is possible to determine whether the vibration of a certain component is responsible for the microphony. A photograph of the equipment in use for testing a valve is shown in Fig. 10.

This method of investigating microphony has the advantages of enabling the responsible component to be located, and immediately showing the effects of modifications. Its main disadvantage is that it is time-consuming.

White-Noise Excitation

The third method for investigating microphony using direct excitation also uses a vibrator. This time the vibrator is not driven at one frequency but by a white-noise signal containing all frequencies. All the parts of the electrode structure of the valve will be driven into resonance simultaneously. Measurement of the r.m.s. value of the interference signal may give an indication of the microphony performance of the valve but by using a frequency-selective amplifier which can 'scan' the complete frequency range investigated, a plot of microphonic voltage against frequency can be obtained. This can be displayed on an oscilloscope or produced as a pen trace.

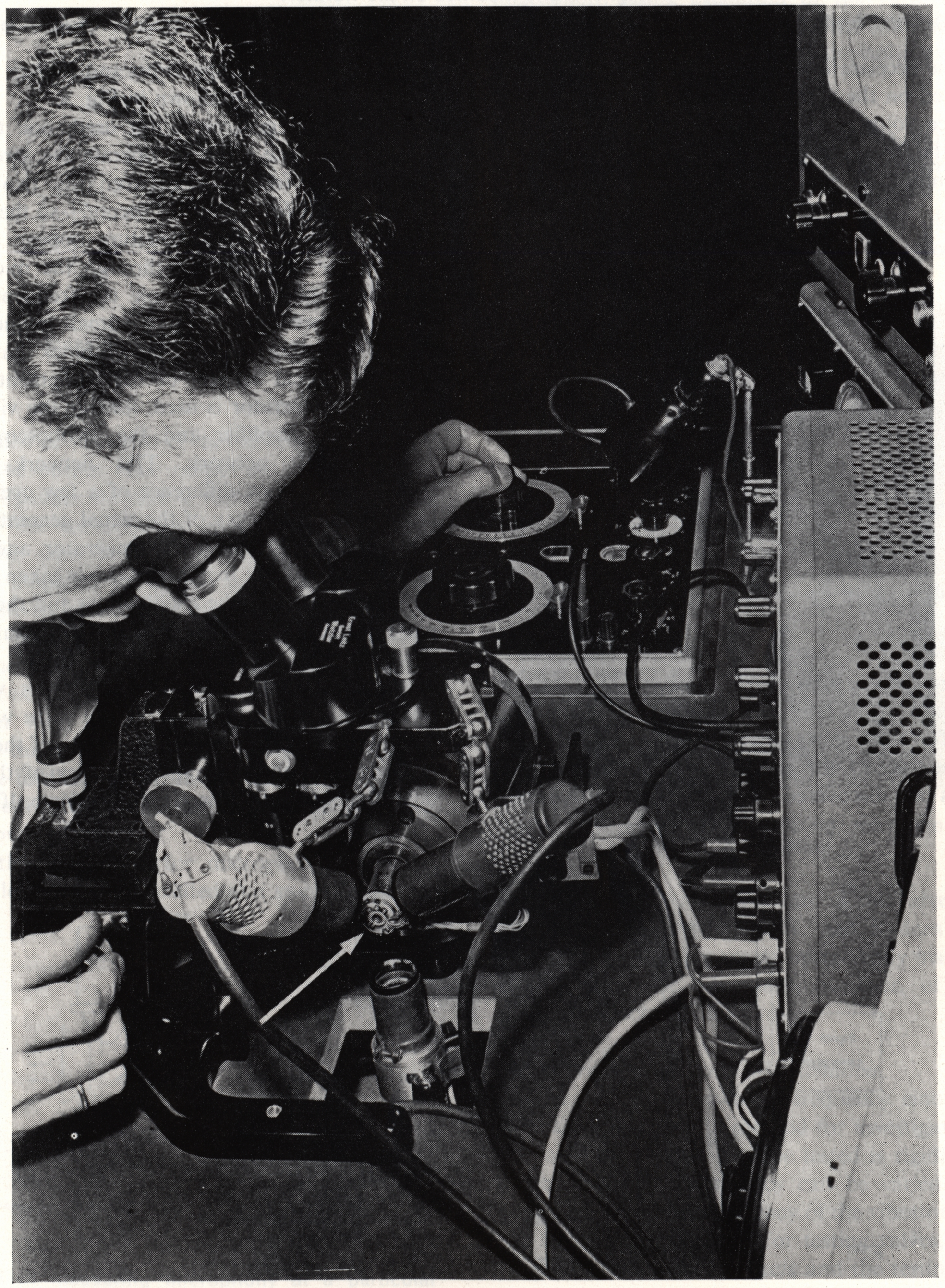


Fig. 10—Photograph of vibrator equipment in use. The arrow points to the valve under test

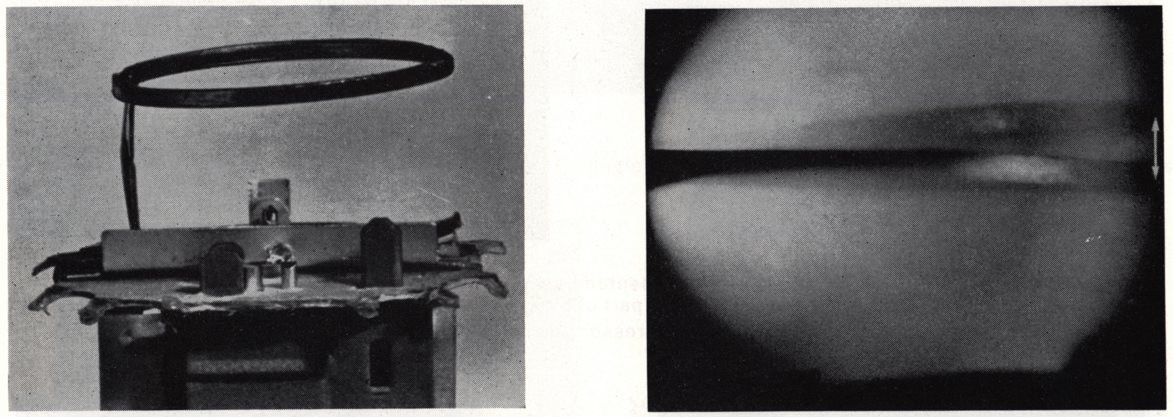
A disadvantage of this method is that the height of the peaks in the display depend on their width. This is because the vertical deflection of the display is proportional to the value of the microphony voltage averaged over the bandwidth of the frequency-selective amplifier. Consequently a peak narrower than the bandwidth of the amplifier will appear shorter than a wider but equally high peak. This has to be remembered when analysing the display.

ILLUSTRATIONS OF MICROPHONY

To end this part of the article, some illustrations of the vibration of electrode structures inside valves will be given. The photographs were taken using the vibrator equipment shown in Fig. 10. A double-exposure technique, each exposure being taken at the extremes of the vibration, enables the amplitude of the vibration to be shown on one photograph.

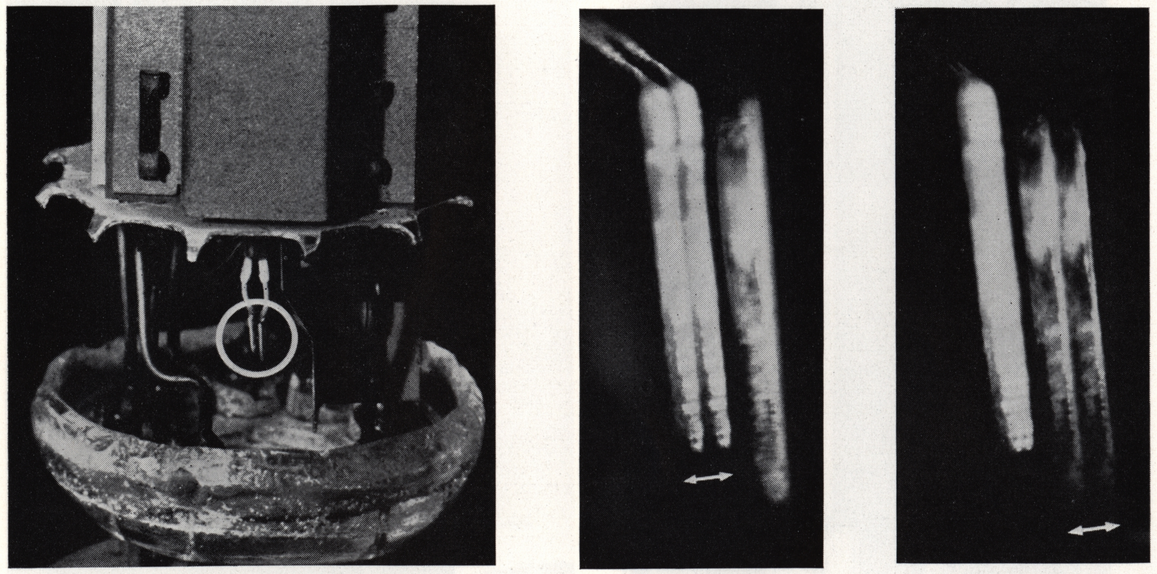
Fig. 11a shows the getter in a valve, fixed at one end only. Fig. 11b shows the vibration of this part at its resonant frequency of 300c/s. Although the getter does not have any electrical effect on the operation of the valve, it may force-excite other electrodes and so adversely affect the operation.

Fig. 12 shows two heater leads of a valve which have different resonant frequencies. Fig. 12a shows the two heater leads, Fig. 12b shows one of the leads vibrating at its resonant frequency of 570c/s, and Fig. 12c shows the other lead vibrating at the resonant frequency of 600c/s.



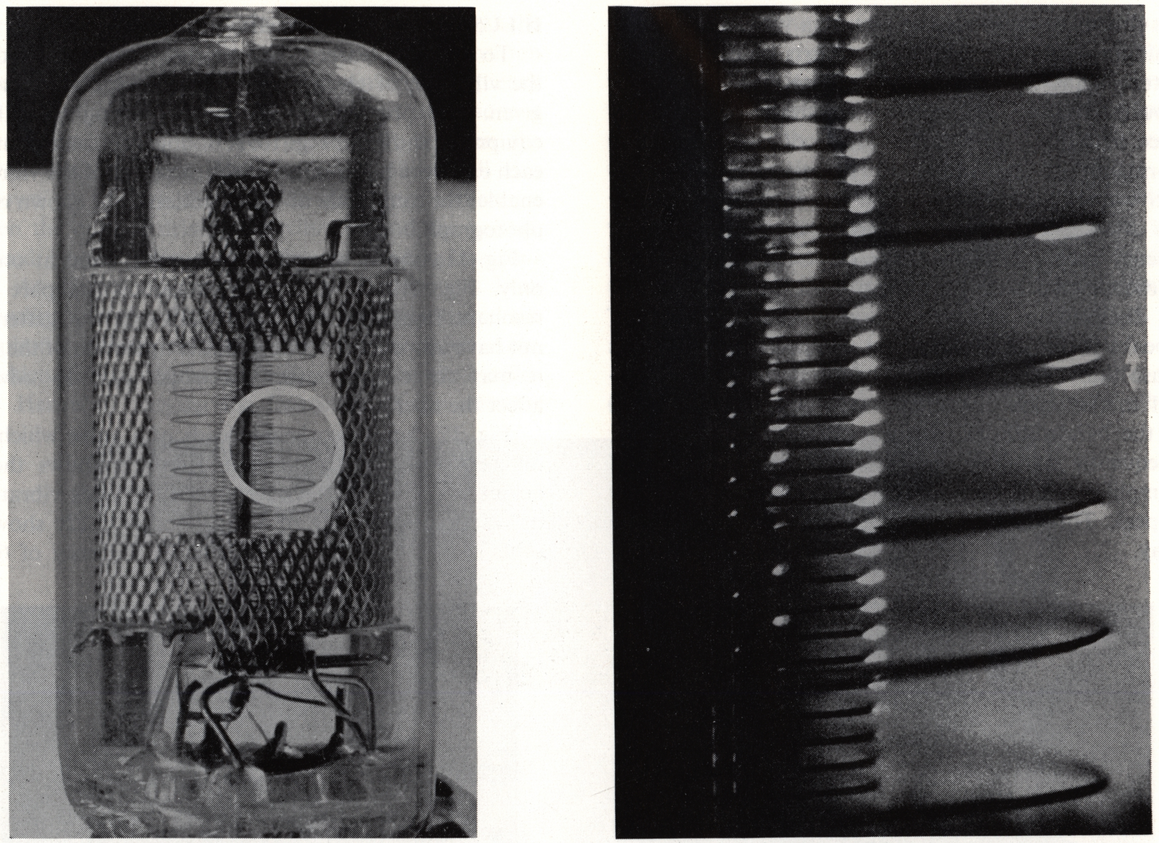
(a) (b)

Fig. 11—(a) Getter of valve fixed at one end only
(b) Getter resonating at 300c/s



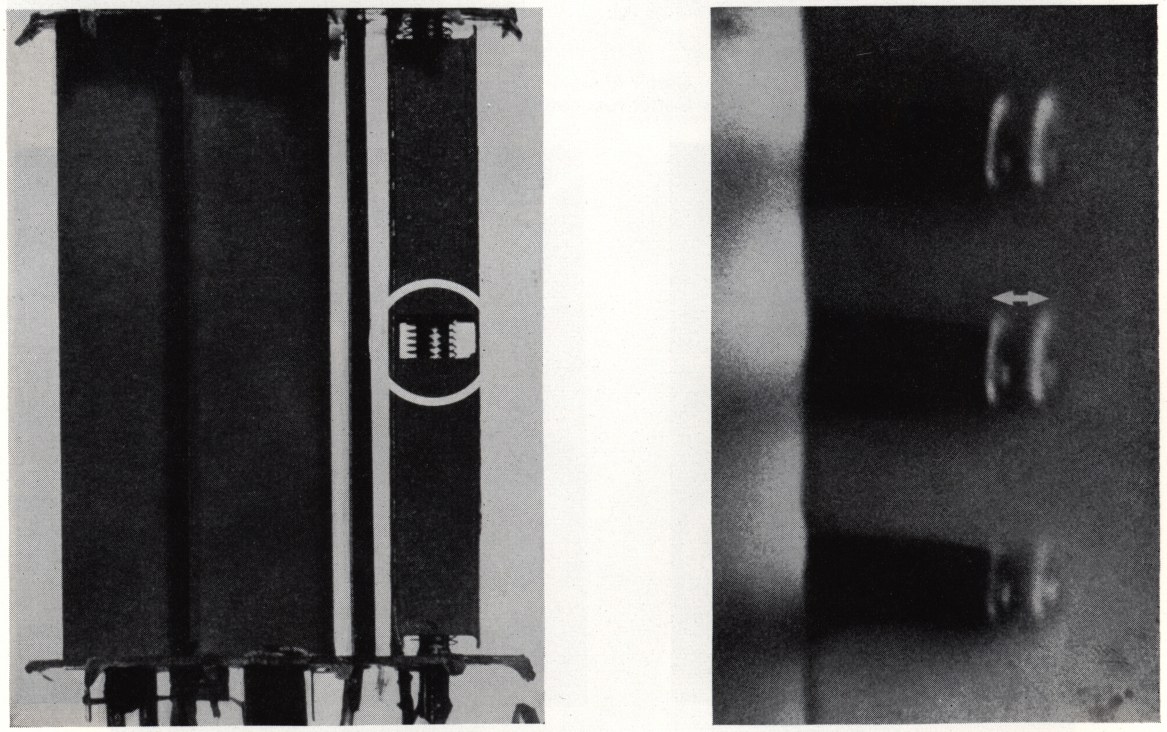
(a) (b) (c)

Fig. 12—(a) Heater leads of valve (in circle)
(b) One heater lead resonating at 570c/s (c) Other heater lead resonating at 600c/s



(a) (b)

Fig. 13—(a) Suppressor grid of pentode seen through opening cut in anode. The circle marks the part of the grid shown in (b)
(b) One turn of suppressor grid resonating at 2100c/s

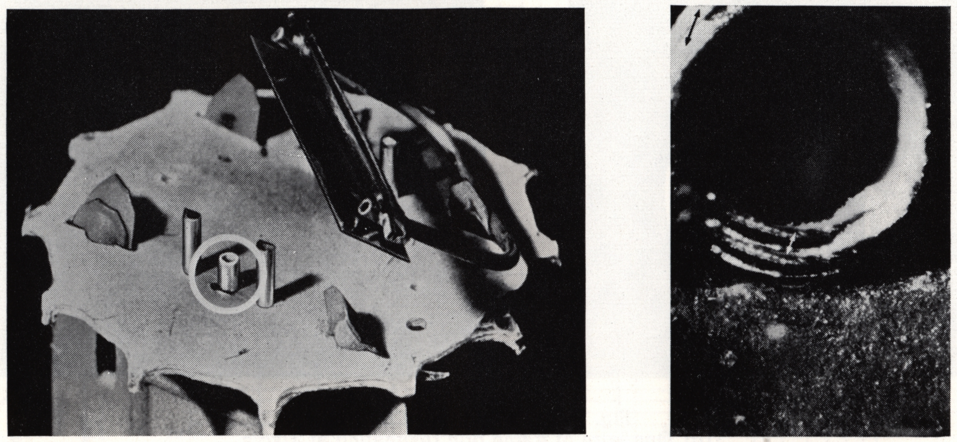


(a) (b)

Fig. 14—(a) Grid of triode section of a triode hexode
(b) Grid resonating at 1900c/s

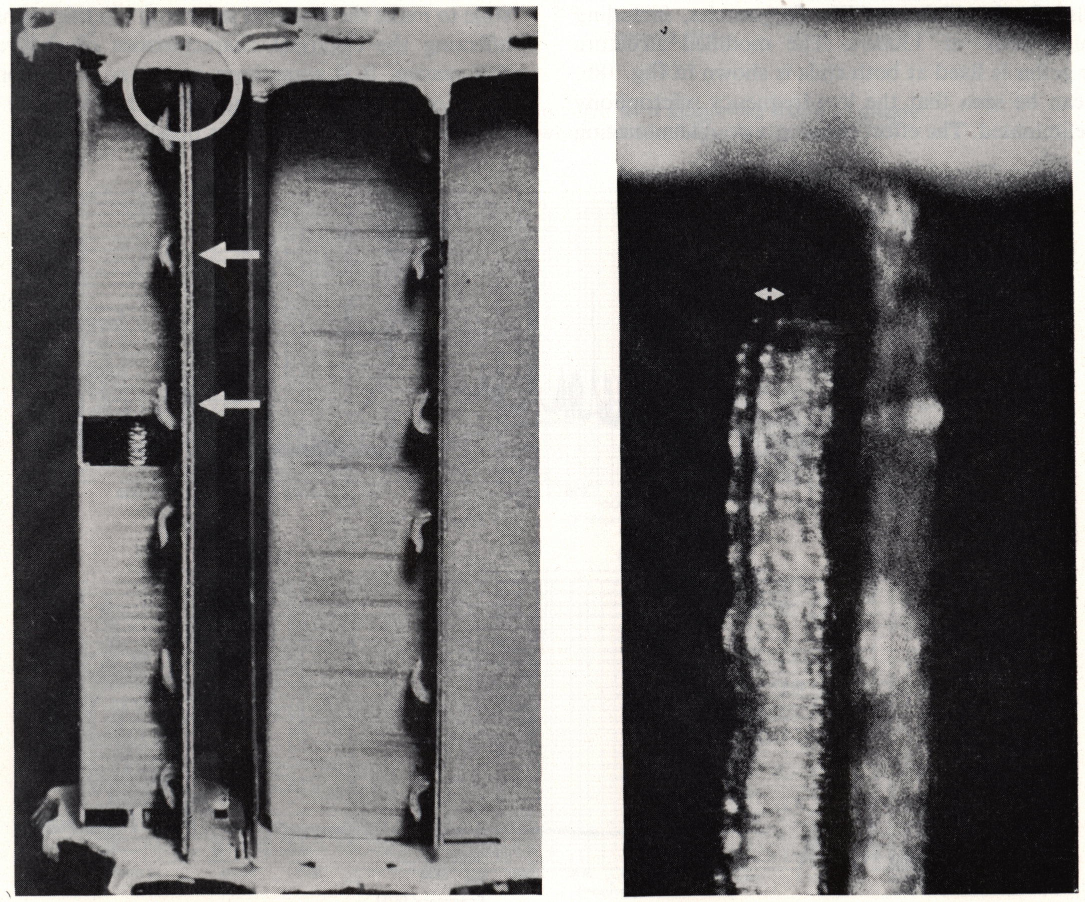
It can be seen that there is little movement of the lead that is not at resonance.

Fig. 13a shows the electrode structure of a pentode valve with the suppressor grid made visible by an opening cut in the anode. In Fig. 13b one of the turns of the grid is at resonance, at 2100c/s, whilst the other turns are at rest. The vibration of the grid of the triode section of a triode-hexode valve is shown in Fig. 14. The resonant frequency this time is 1900c/s. Fig. 15 shows how the slight play between the cathode and the top mica of a valve enables the cathode to vibrate at a frequency of 600c/s. In Fig. 16, insufficiently tight clamping between the two halves of the anode structure has led to movement between them at a frequency of 1300c/s. An example of



(a) (b)

Fig. 15—(a) Top of electrode structure of valve showing (in circle) end of the cathode (b) Cathode resonating at 600c/s

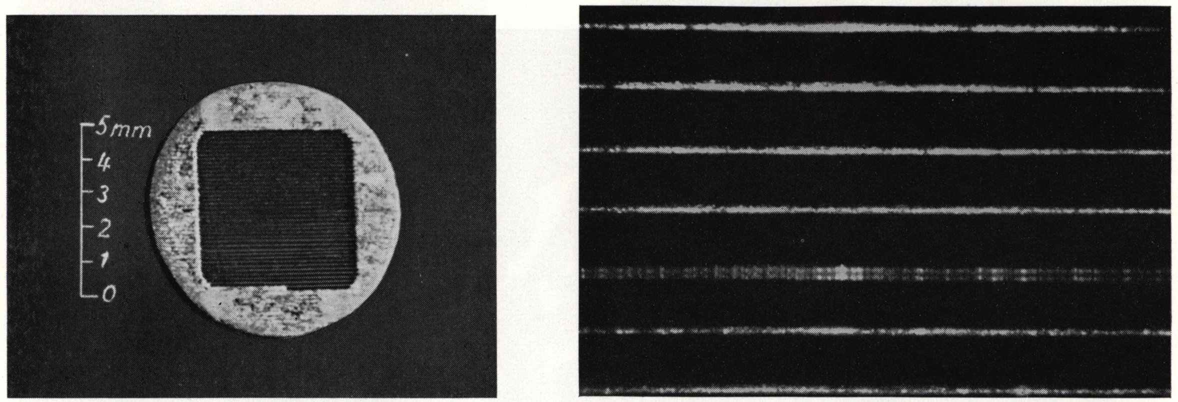


(a) (b)

Fig. 16—(a) Anode structure of valve insufficiently clamped at points marked by arrow
(b) Movement between the parts of the anode at 1300c/s

vibration at higher frequencies is shown in Fig. 17 where one of the wires of a frame grid is resonating at a frequency of 37kc/s.

Pen traces of the output voltage of a valve mounted on the vibrator are very useful for checking the effect of any modifications to the electrode structure of the valve.



(a) (b)

Fig. 17—(a) Frame grid of valve
(b) One wire of frame grid resonating at 37kc/s

In Fig. 18a (derived from a pen trace), the getter of the valve is fixed at one point only and the pen trace shows high levels of microphony at low frequencies, including a very high peak at 1300c/s. The modified structure where the getter is fixed at both ends is shown in Fig. 18b, where it can be seen that the low-frequency microphony has been eliminated. The effect of using a double mounting lug instead of a single lug on the anode structure of a valve is shown in Fig. 19. The double lug ensures better locking of the anode to the mica and so reduces the microphony. The construction of the anode can also affect the microphony, as shown in Fig. 20. The anode shown in Fig. 20a is greatly inferior to that shown in Fig. 20b, as the two pen traces show.

Although the valve manufacturer tries to make valves as free as possible from microphony, it is not always possible to make the required structural modifications without affecting the electrical characteristics of the valve. For this reason, it is sometimes necessary for a compromise to be made. In general, however, the electrode structure of a valve is made as rigid as possible.

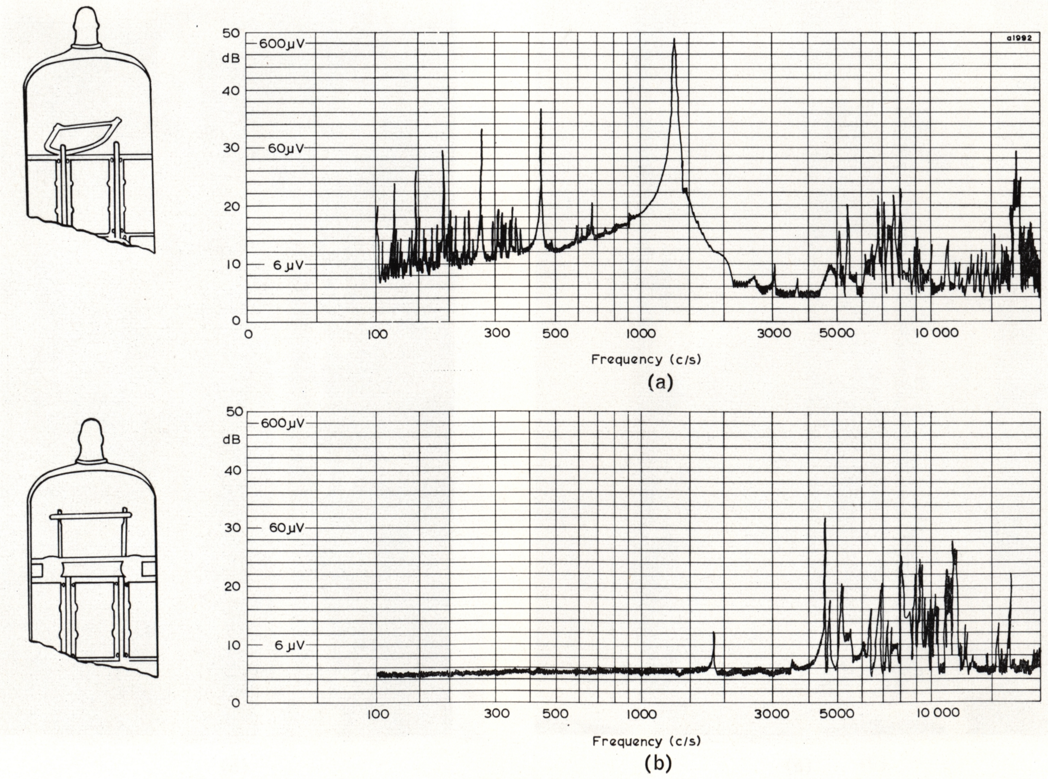


Fig. 18—(a) Pen trace of microphony in valve with getter fixed at one end only
(b) Pen trace of microphony when getter is fixed at both ends

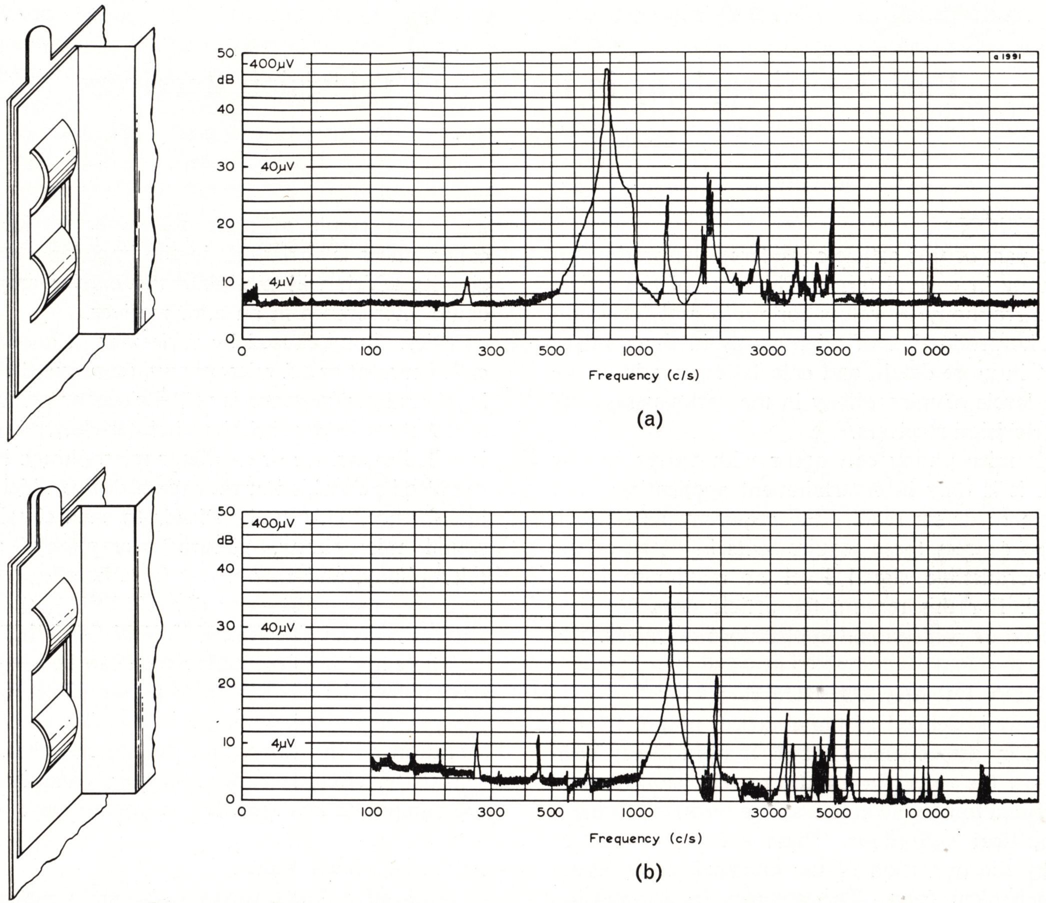


Fig. 19—(a) Pen trace of microphony in valve with single-lug fixing of anode
(b) Pen trace of microphony when double fixing lug is used

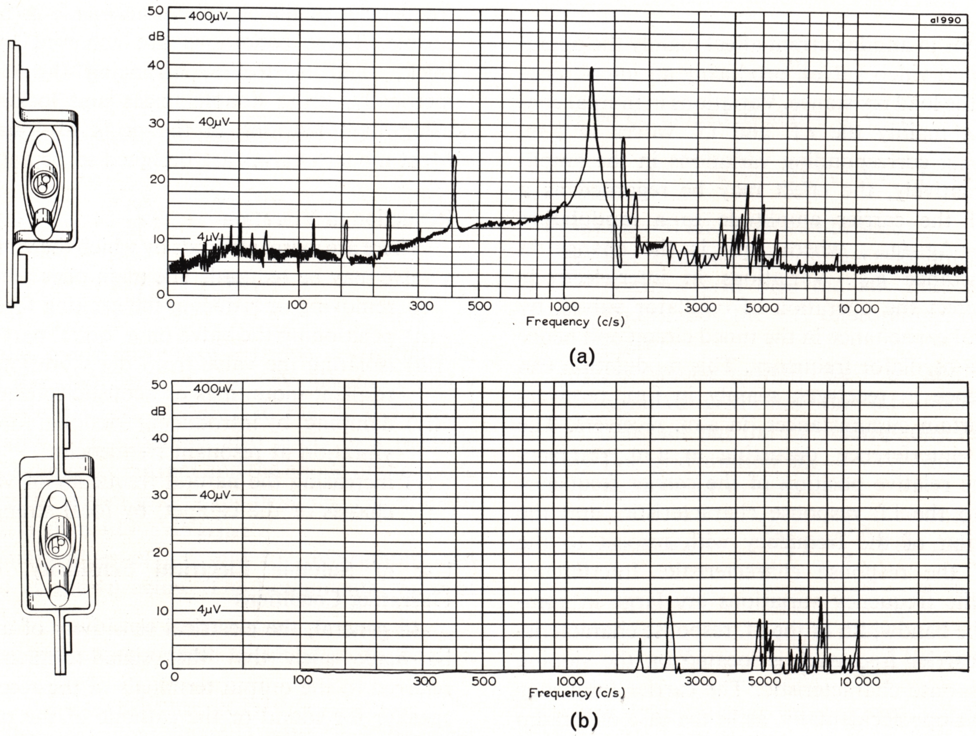


Fig. 20—(a) Pen trace of microphony in valve with assymetrical anode
(b) Pen trace of microphony with symmetrical anode

Part 2—Microphony in Radio and Television Receivers

INTRODUCTION

The first part of this article considered the production of microphony in general terms. This part of the article considers microphony in valves for radio and television receivers. The production of microphony in these valves is described in more detail, and calculations made of the permissible levels of microphony in the various stages of radio and television receivers.

Although microphony can occur with valves in any application, it is only in entertainment applications that it is possible to define readily the requirements of each stage and so express these requirements in terms of the minimum permissible signal level or maximum interference level. For this reason, the article does not deal with industrial or professional applications of valves.

PRODUCTION OF MICROPHONY

Microphony is defined as the electrical output from a valve caused by a mechanical input. It has already been described how the valves in a radio or television receiver can be subjected to varying accelerations through mechanical or acoustical excitations. These excitations can be produced by the operation of the loudspeaker or by an outside mechanical force. The changes in acceleration produce relative motion between the electrodes of the valves and hence, among other effects, cause variations in anode current, mutual conductance, and inter-electrode capacitances.

The variations in anode current affect mainly the operation of audio and video valves, producing an interference voltage across the load resistance. Variations in mutual conductance affect mainly the r.f. and i.f. valves in a.m. receivers, causing corresponding variations in the amplification. Alternatively, the effect may be regarded as a modulation of the carrier amplitude to a modulation depth of $m = \Delta g_m/g_m$, where Δg_m is the change in mutual conductance g_m. Variations in inter-electrode capacitances affect the operation of oscillator valves, by varying the total capacitance in the tuned circuit and hence modulating the oscillator frequency. This modulation can cause interference in receivers, mainly in f.m. receivers but also, in certain circumstances, in a.m. receivers. The possibility of interference occurring in a.m. receivers depends on the relative position of the carrier frequency with respect to the i.f. response characteristic, and the relative variation of the frequency with respect to the bandwidth of the response characteristic. Interference occurs when the frequency variations are large in comparison with the bandwidth of the i.f. response characteristic, or if the carrier frequency is located on the sloping side of the response characteristic. The carrier frequency may be on the slope accidentally, as in the case of a radio which is off tune, or intentionally, as in the case of the vision carrier of a television receiver. A further but less important result of these variations in inter-electrode capacitances is variations in the impedance of the tuned circuits which will modulate the carrier amplitude and cause interference in an a.m. receiver.

Microphony caused by variations in anode current is called current or a.f. microphony, that caused by variations in mutual conductance is called slope or r.f. microphony, and that caused by variations in inter-electrode capacitance is called capacitive or oscillator microphony. Microphony can also be classified by the output device which reveals the interference. This method leads to such classifications as sound microphony, picture microphony, oscilloscope microphony, and so on.

PREVENTION OR REDUCTION OF MICROPHONY

There are three methods which can be used to prevent or to reduce the effects of microphony. These methods are: the use of anti-microphonic valves, the reduction of the vibration on the valve, and the use of suitable electrical sensitivity, circuitry, and operating conditions to make the equipment less sensitive to microphony.

Anti-Microphonic Valves

Anti-microphonic valves are specially manufactured so that when the valve is subjected to vibration, the relative motion of the electrodes is as small as possible. The electrode structure of the valve is firmly fixed and the natural frequencies of the electrodes are made as high as is practicable. These conditions are achieved by using short, thick, rigid electrodes, arranging the electrodes symmetrically, using double micas and locking straps, and using high tensions in the grids of frame-grid valves. Most modern valves are designed to be anti-microphonic.

Reduction in Vibration

There are many methods by which the vibration reaching a valve may be reduced. The main ones are:

(i) removing or reducing the exciting force;
(ii) positioning the valve on a 'quiet' part of the chassis;
(iii) isolating the valve from the vibration, by means of resilient mountings or acoustic screening, etc.;
(iv) damping, by introducing frictional forces to limit the vibration at resonant frequencies;
(v) increasing the natural frequency of vibration of the chassis or the cabinet, by local strengthening.

Use of Suitable Electrical Sensitivity, Circuitry, and Operating Conditions

In general, the electrical sensitivity of a circuit should be made such that the signal-to-microphony ratio, referred to the output terminals of the receiver (the loudspeaker for sound or the cathode of the picture tube for video), is as high as the design considerations allow. As will be shown later, microphony depends to some extent on the electrical sensitivity* of the circuits used and on the operating conditions of the valve (Refs. 1 and 2). As far as the operating conditions are concerned, it is advisable to choose the working point well away from the knee of the I_a/V_g characteristic. The picture and sound microphony of the oscillator section of a frequency-changer valve can be reduced by reducing the extent to which the valve inter-electrode capacitances form part of the oscillator-circuit capacitances, and by increasing the ratio of oscillator capacitance to oscillator frequency.

EXAMPLE

As an example of microphony calculations, the a.f. and r.f. microphony as functions of operating voltage and the relative variation in inter-electrode spacing will be calculated.

For simplicity, a single planar valve system is considered (Fig. 1) with the following assumptions:

(i) the electrodes move uniformly throughout their length in a direction perpendicular to the plane of the electrodes;
(ii) microphony other than that at the fundamental frequency of vibration is negligible;
(iii) the valve operates outside the 'island effect' region.

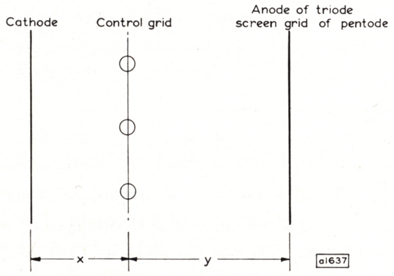


Fig. 1—Single planar valve system

The approximate value of anode current is given by the equations:

$$I_a = \frac{k_1}{x^2}\left(V_g + \frac{V_a}{\mu}\right)^{3/2},$$

$$= \frac{k_1}{x^2}\left(V_g + k_2\frac{V_a}{y}\right)^{3/2}, \qquad \ldots(1)$$

where I_a is the anode current of a triode, V_g is the grid voltage, μ is the amplication factor $[dV_a/dV_g]_{Ia}$, x and y are the electrode spacings as shown in Fig. 1, and k_1 and k_2 are constants. The equations also apply to a pentode if I_a is taken as the space current and the anode of the triode is replaced by the screen grid of the pentode. The amplification factor is then $[dV_{g2}/dV_{g1}]_{Ig2}$.

Fundamental values of a.f. and r.f. microphony can be calculated as shown below. The fundamental value of a.f. microphony is defined by the equation:

$$\Delta V_g = \frac{\Delta I_a}{g_m}, \qquad \ldots(2)$$

where ΔI_a is the change in anode current caused by the vibration and ΔV_g is the equivalent microphonic change in grid voltage corresponding to ΔI_a.

I_a may be expressed as a MacLaurin's series and an expression for ΔI_a derived. Neglecting the second and higher order terms and substituting this expression for ΔI_a in Eq.(2) gives:

$$\Delta V_g = \frac{1}{g_m}\left(\frac{\partial I_a}{\partial x}.\Delta x + \frac{\partial I_a}{\partial y}.\Delta y\right). \qquad \ldots(3)$$

From:

$$g_m = \left[\frac{dI_a}{dV_g}\right]_{V_a},$$

and Eq.(1), it follows that:

$$g_m = \frac{3}{2}\frac{k_1}{x^2}\left(V_g + k_2\frac{V_a}{y}\right)^{1/2}. \qquad \ldots(4)$$

If the expression for g_m of Eq.(4) is substituted in Eq.(3), and the partial differentiation of I_a in terms of Eq.(1) performed, Eq.(3) becomes:

$$\Delta V_g = \left\{\frac{3}{2}\frac{k_1}{x^2}\left(V_g + k_2\frac{V_a}{y}\right)^{1/2}\right\}^{-1}\left\{-\frac{2k_1}{x^3}\left(V_g + k_2\frac{V_a}{y}\right)^{3/2}\Delta x - \frac{3}{2}\frac{k_1}{x^2}\frac{k_2}{y^2}V_a\left(V_g + k_2\frac{V_a}{y}\right)^{1/2}\Delta y\right\}.$$

This simplifies to:

$$\Delta V_g = -\frac{4}{3}\left(V_g + k_2\frac{V_a}{y}\right)\frac{\Delta x}{x} - k_2\frac{V_a}{y}\frac{\Delta y}{y}. \qquad \ldots(5)$$

Similarly the equation for the fundamental value of r.f. microphony:

$$m = \frac{\Delta g_m}{g_m},$$

may be developed as above to yield:

$$m = -2\frac{\Delta x}{x} - \tfrac{1}{2}\left(V_g + k_2\frac{V_a}{y}\right)^{-1}.k_2\frac{V_a}{y}\frac{\Delta y}{y}. \qquad \ldots(6)$$

From Eqs.(5) and (6) it can be seen:

(i) For cathode motion $\Delta y = 0$, and with increasing negative grid bias the a.f. microphony decreases, Eq.(5), but the r.f. microphony remains constant, Eq.(6).
(ii) For anode motion $\Delta x = 0$, and with increasing negative grid bias the a.f. microphony remains constant but the r.f. microphony increases.
(iii) For grid motion $\Delta x = -\Delta y$, and ΔV_g and m can be reduced to zero at:

$$V_g = -\frac{V_a}{\mu}\left(1 - \frac{3}{4}\frac{x}{y}\right) \text{ for } \Delta V_g = 0,$$

and

$$V_g = -\frac{V_a}{\mu}\left(1 - \frac{1}{4}\frac{x}{y}\right) \text{ for } m = 0.$$

In a double planar valve system, the effective values of ΔV_g and m will be reduced by cancellation between one side of the valve and the other. In a valve with perfect

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*Electrical sensitivity with respect to sound microphony is defined as:

$$P_o^{1/2}/v_i,\ P_o^{1/2}/m_i,\ \text{or}\ P_o^{1/2}/f_d.$$

P_o is the loudspeaker output power in watts; v_i is the input voltage to the a.f. valve, m_i is the modulation depth, and f_d the frequency deviation in kc/s, to produce a loudspeaker power P_o.

symmetry, fundamental a.f. and r.f. microphony will be zero but microphony at harmonics may exist. This is not shown in the above equations as only fundamental microphony has been considered. In practice, because of the island effect, both ΔV_g and m may increase with increasing negative grid bias.

SOUND MICROPHONY

The main types of sound microphony are 'feedback howl' (or howlback) and 'clang'.

Feedback Howl

Feedback howl occurs when the initial microphony of a valve is amplified, converted by the loudspeaker into vibrations that are fed to the valve with such a phase relationship and amplitude as to sustain the vibration of the electrodes that produced the initial microphony. It can be seen, therefore, that feedback howl can be prevented, even with a favourable phase relationship, if the microphony from a certain loudspeaker power is made less than the input signal required to produce that same speaker power. The input signal and the microphony should be defined in the same terms – input voltage for an a.f. valve, modulation depth for an i.f. or r.f. valve, and frequency deviation for an oscillator valve.

Since the possibility of feedback howl increases with increasing electrical sensitivity and relative vibration with respect to speaker power, the microphony limits should be expressed in terms of the acceleration caused by a given loudspeaker power and the electrical sensitivity of the valve. The acceleration of a valve in a receiver is proportional to the square root of the loudspeaker output power. In general, the peak acceleration in a receiver is less than 0·2g at 50mW or 1·0g at 1W output power.

If it is assumed that the loudspeaker output power for a television receiver does not exceed 5W and for a radio receiver 12W, and also that these output powers are realised only if the modulation depth with an a.m. receiver is greater than 30% and if the frequency deviation of an f.m. receiver is greater than 15kc/s, the requirements for the prevention of feedback howl can be expressed as below.

A.F. Valves

From the equation for electrical sensitivity, the requirement for the prevention of feedback howl is:

$$\frac{P_o^{1/2}}{v_i} = \frac{(0{\cdot}05)^{1/2}}{v_i'}, \qquad \ldots(7)$$

where v_i is the input voltage for a loudspeaker power of P_o, and v_i' is the input voltage for 0·05W. Eq.(7) can be rearranged as:

$$v_i = v_i'\left(\frac{P_o}{0{\cdot}05}\right)^{1/2}. \qquad \ldots(8)$$

Amplifiers using a.f. valves should therefore be designed, unless otherwise stated in published data, so that the vibration on the valve from a loudspeaker power of 50mW does not exceed the normal value of 0·2g, and the minimum permissible input voltage is defined by Eq.8.

R.F. and I.F. Valves

The requirements for the prevention of feedback howl in television and radio receivers respectively are:

$$\frac{P_o^{1/2}}{m} = \frac{5^{1/2}}{0{\cdot}3} \text{ and } \frac{P_o^{1/2}}{m} = \frac{12^{1/2}}{0{\cdot}3},$$

that is:

$$m = 0{\cdot}3\left(\frac{P_o}{5}\right)^{1/2} \text{ and } m = 0{\cdot}3\left(\frac{P_o}{12}\right)^{1/2}. \qquad \ldots(9)$$

Amplitude-modulation receivers incorporating r.f. and i.f. valves should therefore be designed so that the vibration on the valve does not exceed the normal value, and that at maximum sensitivity a speaker power P_o cannot be produced by signals having a modulation depth less than that defined in Eq.(9).

Oscillator Valves

The requirements for the prevention of feedback howl from the oscillator valve in f.m. receivers are:

$$\frac{P_o^{1/2}}{f_d} = \frac{5^{1/2}}{15} \text{ and } \frac{P_o^{1/2}}{f_d} = \frac{12^{1/2}}{15},$$

for television and radio receivers respectively; that is:

$$f_d = 15\left(\frac{P_o}{5}\right)^{1/2} \text{ and } f_d = 15\left(\frac{P_o}{12}\right)^{1/2}. \qquad \ldots(10)$$

Frequency-modulation receivers should therefore be designed so that the vibration on the oscillator valve does not exceed the normal value, and that at full sensitivity a speaker power P_o cannot be produced by signals with a frequency deviation less than that defined by Eq.(10).

Clang

Clang is microphony caused by shock excitation; for example, by the operation of a motor or a switch, or by tapping the chassis or cabinet of a receiver. The effect of this type of microphony on the ear depends on several factors. The most important are: the amplitude, frequency and duration of the clang; the comparison of the clang with the mechanical noise coming directly from the shock source; and the comparison of the clang with the useful signal.

These factors lead to the following limitations for clang microphony:

(i) clang should not produce a more objectionable impression than the mechanical noise coming directly from the shock source;

(ii) The signal-to-clang ratio should be greater than 34dB for continuous excitation (for example, caused by a motor), or greater than 24dB for intermittent excitation (for example, caused by a switch).

Example

As a second example of microphony calculations, the determination of the maximum permissible acceleration for an EF86 in a tape recorder is shown. The valve is assumed to be a pre-amplifier with a voltage from the replay head of 0·5mV for a loudspeaker power of 50mW.

If a safety factor of 2 is used, the maximum permissible

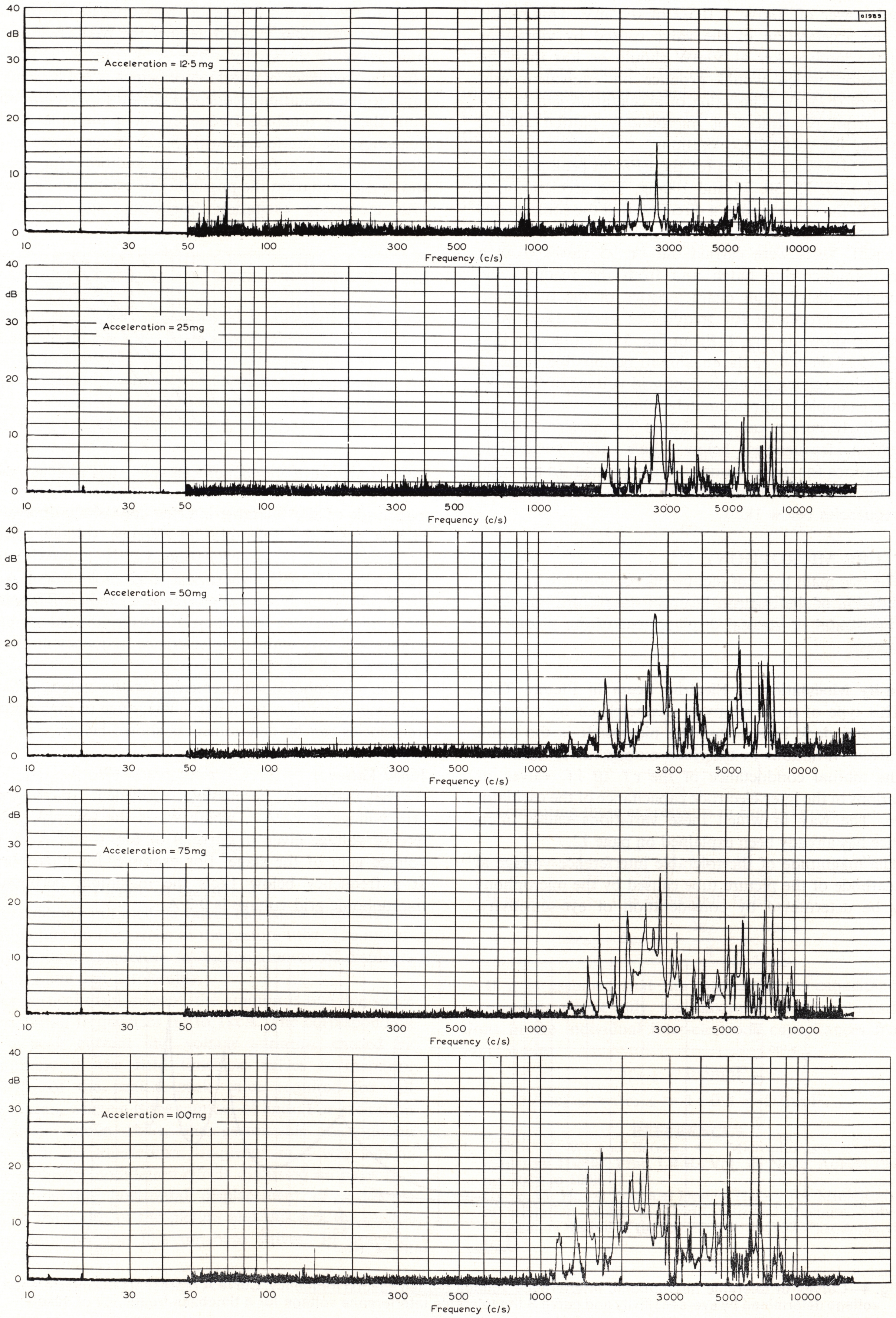

Fig. 2 (right)—Pen traces of microphony of an EF86 plotted against frequency for different peak acceleration levels. 0dB corresponds to an equivalent microphonic grid voltage of 6·67µV

equivalent microphonic grid voltage for a loudspeaker output power of 50mW is 0·5/2 or 0·25mV. This means that the acceleration of the valve caused by vibrations from the loudspeaker at 50mW should be less than the acceleration which would produce an equivalent grid voltage of 0·25mV. The value of acceleration can be obtained from pen traces of microphony plotted against frequency for various accelerations.

Fig. 2 shows such traces for the EF86 at peak accelerations of 12·5, 25, 50, 75 and 100mg. The acceleration is applied to the valve in such a direction that the components of the acceleration along the three main axes of the valve are equal. Tests were carried out on 75 valves from a large manufacturing batch and from the readings the cumulative distribution characteristics of the maximum observed microphony for frequencies above 1kc/s were drawn, as shown in Fig. 3. From Fig. 3 it can be seen that for 99% of the valves the maximum permissible value of acceleration (giving an equivalent grid voltage of 0·25mV) for the example is approximately 20mg.

Similarly curves can be plotted for accelerations at excitation frequencies below 1kc/s and permissible accelerations for these frequencies determined. In practice, however, no microphony trouble occurs with the EF86 at frequencies below 1kc/s.

PICTURE MICROPHONY

The two main types of picture microphony are intensity modulation of the electron beam in the picture tube, and the horizontal or vertical displacement of the picture on the screen.

Intensity Modulation of Electron Beam in Picture Tube

This type of microphony is characterised by alternate bright and dark bars across the picture. It is the result of current variations in the video output valve, variations in the mutual conductance of the r.f. or i.f. valves, or variations in the inter-electrode capacitance of the oscillator valve. The way in which these variations cause interference has already been explained on page 12.

The magnitude of the tolerable interference voltage at the cathode of the picture tube caused by the microphony may be determined by consideration of eye-sensitivity. A curve of the tolerable interference voltage for constant excitation at a given frequency, measured at the cathode of the picture tube, plotted against the frequency of the interference is shown by curve A in Fig. 4. However, when the frequency of interference (which is not necessarily equal to the excitation frequency) deviates by an amount f from the line timebase frequency, or a multiple of this frequency, the interference will appear on the screen of the picture tube at a frequency f. In addition, for speech and music excitation the possibility of excitation decreases with increasing excitation frequency, and hence the magnitude of the tolerable interference voltage increases. In this case, the tolerable interference voltage caused by speech and music excitation v_m is approximately equal to $v_{tol}(f_{exc})^{1/2}$, where v_{tol} is given by curve A of Fig. 4 and f_{exc} is the excitation frequency in kc/s. Therefore for a line timebase frequency of 10kc/s, the interference voltage will vary with the interference frequency as shown in curve B of Fig. 4.

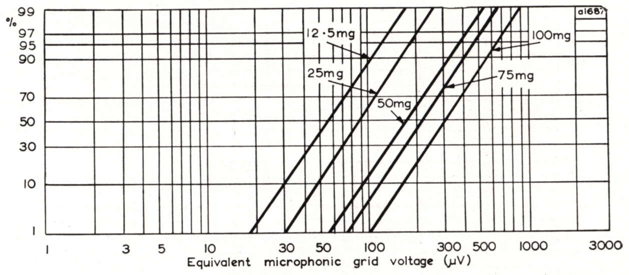

Fig. 3—Cumulative distribution characteristics of EF86 valves

The microphony requirements for the video, r.f., i.f. and oscillator valves for a particular range of microphony frequencies can be determined as shown below.

Video Valves

The maximum permissible equivalent microphonic grid voltage is v_m/A_v, where v_m is the tolerable interference voltage given in curve B of Fig. 4 and A_v is the voltage gain of the stage. For example, for microphony at a frequency below 1kc/s the maximum permissible microphonic grid voltage is $0{\cdot}2/A_v$.

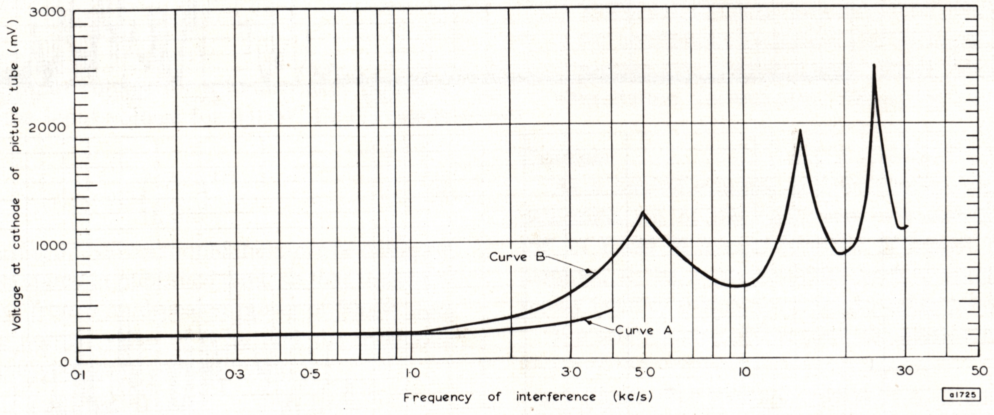

Fig. 4—Curves of tolerable microphonic interference voltage plotted against frequency. Curve A represents the interference voltage determined by eye-sensitivity and curve B the tolerable interference voltage for a timebase frequency of 10kc/s

R.F. and I.F. Valves

It is assumed that the peak-to-peak video signal at the cathode of the picture tube is 70V, representing 65% of the peak-to-peak carrier voltage. For linear amplification and detection, the permissible relative variation in slope or microphonic modulation depth is:

$$m = v_m \times \frac{2\sqrt{2}}{70/0{\cdot}65}.$$

For microphonic frequencies below 1kc/s:

$$m = 0{\cdot}2 \times \frac{2\sqrt{2}}{70/0{\cdot}65},$$
$$\simeq 0{\cdot}5\%.$$

Oscillator Valves

The video i.f. carrier is located in the middle of the sloping side of the video i.f. response characteristic. The width of this sloping side is approximately 1500kc/s so that a frequency deviation of ±750kc/s will correspond to 100% modulation depth. The permissible modulation depth of 0·5% calculated above corresponds to a frequency deviation of ±3·75kc/s.

The relationship between the frequency variation Δf_o and the inter-electrode capacitance variation Δc_o caused by vibration of the oscillator valve is given by:

$$\frac{\Delta f_o}{\Delta c_o} = \frac{f_o}{2c_o}, \qquad \ldots(11)$$

where f_o is the oscillator frequency and c_o the effective oscillator capacitance. Eq.(11) can be obtained by differentiating f_o with respect to c_o in the equation:

$$f_o = \frac{1}{2\pi(L_o c_o)^{1/2}}.$$

For a frequency deviation of ±3·75kc/s, the corresponding variation in oscillator capacitance is given by Eq.(11) as:

$$\Delta c_o = \pm\frac{2c_o}{f_o} \times 3{\cdot}75 \times 10^3. \qquad \ldots(12)$$

A typical value for c_o is 10pF, and for channel 13 f_o is 250Mc/s. Hence:

$$\Delta c_o = 0{\cdot}3\text{mpF}.$$

Displacement of Picture on the Screen

Microphony in timebase valves can cause displacement of the picture on the screen of the picture tube. The microphony limits are therefore set by the amount of displacement of a line that is tolerable.

For vertical displacement, the amount is one line space for excitation frequencies up to 1kc/s. The corresponding peak-to-peak current and voltage variation caused by microphony in the field timebase valves is therefore (100/n)% of the peak-to-peak value required for full vertical scan, where n is the number of lines. For excitation frequencies above 1kc/s, the permissible microphony varies with the interference frequency in a way similar to the intensity modulation effects discussed on page 16.

The tolerable horizontal displacement of a line is 0·2% of the width of the picture and this sets the microphony limit for the line timebase valves. In practice, this corresponds to a horizontal displacement of 1mm on a 19 inch tube.

Example

For a third example, the maximum permissible variation of inter-electrode capacitance will be determined for a Colpitts oscillator forming the oscillator section of the frequency changer in a television receiver. Only interference frequencies up to 1kc/s will be considered. The oscillator frequency is 250Mc/s, and the oscillator capacitance is 10pF. As calculated previously the maximum permissible value of Δc_o is 0·3mpF.

The capacitance c_o is formed from several capacitances and may be expressed as:

$$c_o = C + c_{a-g} + \frac{c_{g-k}c_{a-k}}{c_{g-k}+c_{a-k}}, \qquad \ldots(13)$$

where C is the stray capacitance of the circuit; c_{a-g} is the anode-to-grid capacitance, c_{g-k} is the grid-to-cathode capacitance, and c_{a-k} is the anode-to-cathode capacitance of the valve.

If C is constant, the incremental oscillator capacitance caused by changes in the various inter-electrode capacitances can be determined by partial differentiation of c_o. This yields:

$$\Delta c_o = \frac{\partial c_o}{\partial c_{a-g}}\Delta c_{a-g} + \frac{\partial c_o}{\partial c_{g-k}}\Delta c_{g-k} + \frac{\partial c_o}{\partial c_{a-k}}\Delta c_{a-k} \qquad \ldots(14)$$

If Eq.(13) is differentiated in turn with respect to c_{a-g}, c_{g-k}, and c_{a-k} the following three expressions are obtained:

$$\frac{\partial c_o}{\partial c_{a-g}} = 1,$$
$$\frac{\partial c_o}{\partial c_{g-k}} = \left(\frac{c_{a-k}}{c_{g-k}+c_{a-k}}\right)^2,$$
$$\frac{\partial c_o}{\partial c_{a-k}} = \left(\frac{c_{g-k}}{c_{g-k}+c_{a-k}}\right)^2.$$

Substituting these expressions in Eq.(14) gives:

$$\Delta c_o = \Delta c_{a-g} + \left(\frac{c_{a-k}}{c_{g-k}+c_{a-k}}\right)^2 \Delta c_{g-k} + \left(\frac{c_{g-k}}{c_{g-k}+c_{a-k}}\right)^2 \Delta c_{a-k}.$$

If $c_{g-k} = c_{a-k}$, this expression reduces to:

$$\Delta c_o = \Delta c_{a-g} + \frac{\Delta c_{g-k}}{4} + \frac{\Delta c_{a-k}}{4}. \qquad \ldots(15)$$

If it is assumed that the inter-electrode capacitances vary one at a time only, the maximum permissible variations in the individual inter-electrode capacitances from Eq.(15) are:

$$\Delta c_{a-g} = 0{\cdot}3\text{mpF},$$
$$\Delta c_{g-k} = \Delta c_{a-k} = 4 \times 0{\cdot}3 = 1{\cdot}3\text{mpF}.$$

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Read-out and On-Off Circuits using a BPY10 Silicon Photovoltaic Cell

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Operation of the Mullard BPY10 silicon photovoltaic cell in series and parallel configurations is examined, with particular reference to tape and card reading, edge detection, and other on-off applications. The preferable parallel configuration is analysed, and design and setting-up procedures are given, with recommendations concerning amplifier circuits to follow the detector stage.

INTRODUCTION

The Mullard BPY10 photovoltaic cell is a silicon p-n junction, moulded in resin for chemical protection and physical strength, with an accurately masked window that allows access to light over a defined area 1·68 × 1·68mm (0·066 × 0·066in). The rectangular shape, with lead-outs at one end (as shown in the photograph), is designed to facilitate stacking for in-line read-out. The dimensions, 0·086 × 0·088 × 0·25in, allow a row of six devices to be arranged under standard 11/16in tape with five digit holes and a location hole. The width of the cell is controlled within ±0·001in.

The device is intended for on-off light-detection applications, such as tape and card reading and edge detection, where detection of light over a small area is required. It may also be used for revolution counting, modulated light beam detection, and optical sound-track pickup. The spectral response peaks at 0·8μm, so that the device is sensitive to a tungsten lamp.

CHARACTERISTICS

Under dark conditions the p-n junction acts as a normal silicon diode junction with high ohmic leakage when a voltage is applied to it (Fig. 1). When energy quanta in the spectral response band impinge on it, hole-electron pairs are created, and those carriers of the appropriate polarity in the depletion layer are swept over the junction potential barrier as free carriers.

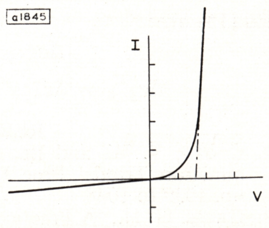
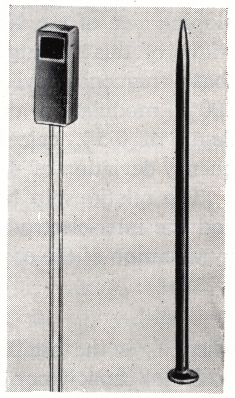


Fig. 1—Dark characteristic of diode

BPY10 and a one-inch pin



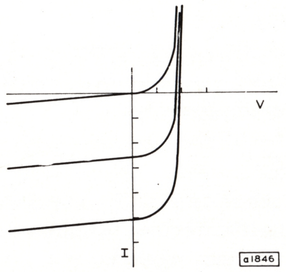


Fig. 2—Effect of incident light

If a low-resistance circuit is connected to the device, these carriers appear as a current in the external circuit, in the reverse direction of the diode. If a high-resistance circuit is connected to the device, the excess free carriers lower the internal barrier field at the junction, causing a potential at the terminals in the forward direction of the diode. If the circuit has a resistance intermediate between these limiting cases, the device acts as a variable generator with non-linear resistance, giving a combination of forward voltage and reverse current as shown in Fig. 2. This characteristic may be considered as consisting of the diode characteristic of Fig. 1 together with a current, proportional to the incident energy, added in the reverse direction. Reversal of the sign of the current axis gives the more usual presentation of these curves, as shown in Fig. 3.

As with all semiconductor diodes, the characteristic is temperature-dependent. The theoretical diode characteristic is given by

$$I = I_s(\exp(qV/KT)-1)$$

where I_s is the saturation reverse current and is exponential with temperature. A leakage current, which is approximately ohmic, occurs across the junction, and this also varies exponentially with temperature. Adding the light current $-c\phi$ (where ϕ is the illumination and c the sensitivity) shifts the curve negatively, giving the curves of Fig. 3, where

$$I = I_s[\exp(qV/KT)-1]+V/f(T)-c\phi.$$

Measurements on the device show that the reverse leakage current measured at −1V bias and dark conditions increases with temperature to 70μA at 100°C, the published maximum rated temperature. The short-circuit sensitivity, when the leakage current may be ignored, is also subject to a small variation with temperature of the order of 0·2% per degC.

CIRCUIT CONSIDERATIONS

The BPY10, being photovoltaic, requires no bias for normal operation, and as the leakage is small the cell may be used with direct coupling into an amplifier. It is, however, desirable to use silicon transistors in the amplifier, so that the transistor leakage current shall not detract from the sensitivity of the cell. In this case, it is necessary to add a bias to the voltage generated by the cell, the total voltage being sufficient to overcome the high base-emitter voltage of the transistor. There are two ways of providing this voltage bias.

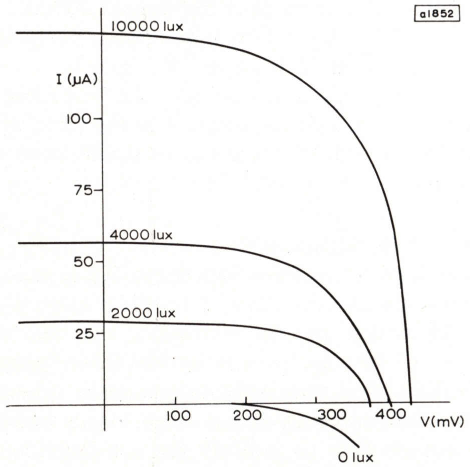


Fig. 3—Typical voltage/current characteristic of BPY10 at $T_{amb} = 20°C$, using a lamp of colour temperature 2700°K

The curves of Fig. 3 show that if the diode is back biased, the characteristics tend towards a constant current state, and the voltage across the device may be increased beyond the V_{be} of the transistor (0·4 to 0·7V) without appreciably affecting the current. If the cell is connected in the reverse direction across the base-emitter diode of the transistor, and the two are fed from a constant current source (Fig. 4) then the transistor will receive a current I_b which is the difference between the current source I_g and the cell current I_d which is proportional to the incident light. The diode voltage will adjust itself to the base-emitter voltage without affecting the diode current. The transistor will also protect the cell, as the current source may be chosen so that when all the current feeds into the transistor, the voltage generated does not exceed the maximum diode reverse voltage.

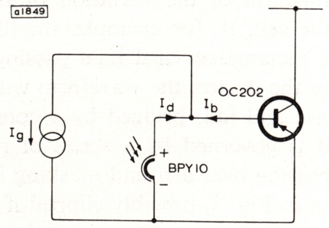


Fig. 4—Parallel configuration

The second method of biasing is to connect the device in series with the transistor base, and to add a bias (Fig. 5) so that the bias provides the base-emitter voltage, and the diode current can force the transistor into conduction. By making the source resistance of the bias supply low, and adjusting the bias level, it is possible to keep the diode voltage small, which keeps the temperature-dependent components of the current small.

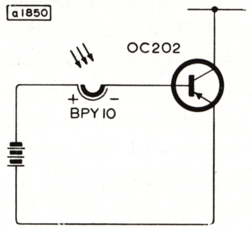


Fig. 5—Series configuration

PARALLEL CONFIGURATION

A circuit employing the device in the parallel configuration is shown in Fig. 6.

As already mentioned, this configuration has the advantage that the cell resistance is high and the cell can be used as a current switch. It has the disadvantage of working at the base-emitter voltage, which may be as much as 0·7V reverse bias with attendant high leakage currents. A sample calculation is shown overleaf.

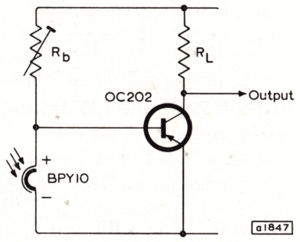


Fig. 6—Practical parallel circuit

To ensure that the circuit is stable the circuit must allow for the following variations:

(i) Drift of V_{cc} power supply; for example, 12V±6%.
(ii) Variation of leakage current up to 30μA at 1V bias and 75°C, which is equivalent to 20μA at 700mV bias, the approximate working point.
(iii) Variation of transistor characteristics. Thus, the current amplification factor h_{FE} may be between 25 and 85 when an OC202 transistor is used.
(iv) Variation in the relative amounts of light falling on the cell in the ON and OFF conditions.
(v) Variation of the short-circuit sensitivity of the device—approximately ±10%.

The light falling on the cell in the OFF condition is determined by the light transmission of the tape or card that is being read (paper tape may transmit as much as 50% of the light falling on it; card transmits between 5% and 20%) and also by the intensity of the light source. As the light output of a lamp varies approximately cubically with applied voltage, it is advisable to power the lamp from a stabilised supply. A typical variation of ±1% in the supply will then give ±3% variation in the light intensity.

Calculation

If I_l and I_s are the leakage and light current of the diode and I_b the base current, then

$$I_b = \frac{V_{cc}-V_{be}}{R_b} - I_l - I_s + I_{CBO} \qquad \ldots(1)$$

The collector leakage current I_{CBO} of the OC202 is small (1μA max) compared with I_s (20μA at 75°C) and may be ignored. Variation in V_{be} is small compared with V_{cc} and may also be ignored.

In the OFF state we may set R_b just to cut off the first transistor of the amplifier at the most adverse limits of the parameters. Then V_{cc} becomes $V_{cc}+6\%$, R_b becomes $R_b-1\%$, and from I_l must be subtracted the variation in illumination and variation of sensitivity. Thus

$$\left(\frac{V_{cc}-V_{be}}{R_b} + 7\%\right) - 0 - (I_l - 3\% - 10\%) = 0$$

and

$$\frac{V_{cc}-V_{be}}{R_b} = \frac{I_l}{1{\cdot}23}. \qquad \ldots(2)$$

In the ON state we wish to bottom the transistor, which requires a base current of $(V_{cc}/R_L)(1/h_{FE})$. Again substituting into Eq(1), using the most adverse tolerance limits, and defining k as the ratio of the illumination in the ON and OFF states, we obtain

$$\frac{V_{cc}}{R_L}\frac{1}{h_{FE}} = \left(\frac{V_{cc}-V_{be}}{R_b} - 7\%\right) - (kI_l + 3\% + 10\%) - \\ -20\mu\text{A (at 75°C)} \qquad \ldots(3)$$

From Eqs(2) and (3), taking V_{cc} as −12V, and h_{FE} as 25, we obtain

$$I_l > \frac{20+(0{\cdot}48/R_L)}{0{\cdot}76-1{\cdot}13k}\mu\text{A} \qquad \ldots(4)$$

where R_L is in MΩ and I_l in μA.

If R_L is at least 470kΩ, the fraction $0{\cdot}48/R_L$ may be ignored.

Examples

(i) For card reading, where k is approximately 0·15,

$$I_l = 36\mu\text{A}.$$

As the minimum sensitivity is 15μA at an illumination of 2000 lux, the minimum light intensity would in this case be 36/15×2000 = 4800 lux, which is easily attainable.

(ii) For tape reading, using paper tape, k may be as high as 0·5. Then from Eq(4)

$$I_l > \frac{20}{0{\cdot}2} = 100\mu\text{A at 75°C}$$

The minimum illumination is 100/15×2000 = 13 400 lux.

More Critical Applications

If it is desired to work at higher temperatures and lower illumination levels than those analysed above, the leakage current will necessitate individual setting up of circuits. This may be avoided by a modification to the circuit.

The leakage current $I_s[\exp(qV/KT)-1]+V/f\,(T)$ is voltage-dependent. Therefore if a bias is added to the device to overcome the transistor base-emitter voltage, the device works at low voltage and the leakage current is considerably reduced. Fig. 7 shows the application of a bias voltage, stabilised by a silicon diode, in series with the cell. The cell is now loaded by the base-emitter diode of the transistor, in series with the external diode, which together present a load of the order of 200Ω. With a maximum signal current of 200μA, only 40mV is developed across the cell, together with any difference between the forward voltages of the diode and the transistor base-emitter diode. The total does not exceed 200mV, which limits the leakage current to approximately one-fifth of its value at 1V (that is, 14μA at 100°C).

As the diode characteristic and the transistor input characteristic vary with temperature in the same way, the diode will also provide temperature stabilisation of the working point.

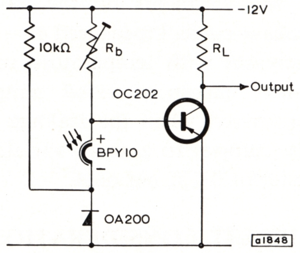


Fig. 7—Parallel circuit with bias

SERIES CONFIGURATION

Under *Circuit Considerations* (page 19) a series configuration was suggested, in which the diode current is used directly to switch on the transistor, the base-emitter voltage being overcome by a series bias (Fig. 5). As there is no parallel path in this configuration, the leakage current flows into the transistor, and it is no longer possible to use the cut-off state to indicate the OFF condition. The circuit may be adjusted so that the transistor just bottoms in the ON condition by adjustment of the collector load. This, however, makes the operation of the circuit dependent on h_{FE}, which is temperature-dependent, and on the input resistance of the following amplifier, which may also vary. The series configuration is therefore inferior to the parallel configuration, and a detailed analysis will not be given.

FOLLOWING CIRCUITS

The waveform from either of the above circuits is a function of the variation of the illumination falling on the cell. If, for example, the illumination is governed by a rectangular card hole passing over the square window of the device, the waveform will consist of a nearly linear rise and fall, bridged by a constant voltage. If, however, it is governed by a series of round holes in paper tape passing over a round masking hole, the waveform will be as in Fig. 8, possibly clipped if a high-sensitivity diode or low-transparency tape is used.

If square pulses are required, a switching circuit must be used after the first transistor to define the ON and OFF states. The most suitable circuit for this purpose is a Schmitt trigger circuit (Ref. 1), an example of which is given in Fig. 9. As the circuit is drawn, it gives positive pulses. The circuit is capable of reading pulses from a paper tape with k = 0·3 and an illumination of 5000 lux with any cell, provided that the ambient temperature does not exceed 75°C.

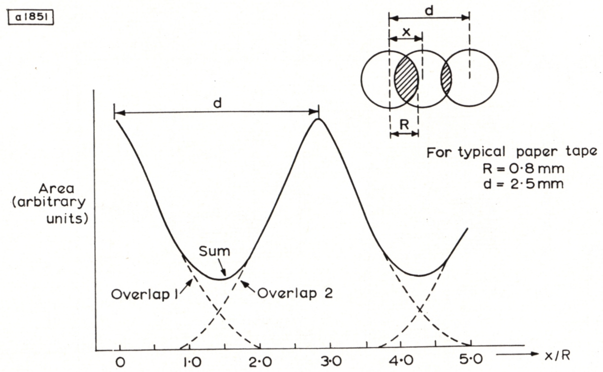


Fig. 8—Area of overlap of a round hole passing over a row of similar holes

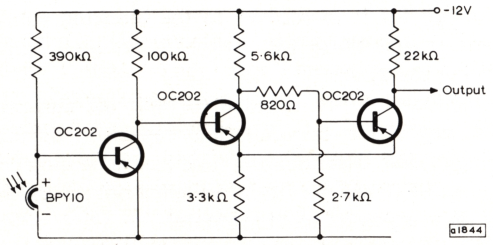


Fig. 9—Schmitt trigger circuit

CONCLUSIONS

The BPY10 cell is suitable for tape and card reading, and similar on-off light detection applications, in an ambient temperature of up to 75°C, with illumination levels in the ON state in the range 2000 to 10 000 lux, provided that the illumination in the OFF state does not exceed 30% of that in the ON state. The majority of tape and card readers should fall well into this group.

If it is desired to work at temperatures above 75°C, with an illumination less than 10 000 lux, it may be necessary to use low-transparency tape, or to adjust each circuit individually.

REFERENCES

1. NEWELL, A. F. and TOURTEL, P. A. 'Transistor Backlash Circuits'. **Mullard Technical Communications,** Vol. 6, No. 51, Sept. 1961, pp. 22 to 29.

APPENDIX

Setting Up the Circuit

The sensitivity of the device may have any value in the range 15μA to 50μA at 2000 lux. If the circuit is designed to accept the lowest sensitivity cells, and a high-sensitivity cell is used, the output in the OFF state, using a card or tape with an ON/OFF light transmission ratio of less than 10, will be of the same order as that in the ON state. In the majority of cases it is therefore necessary to match the circuit to each individual cell by means of a set-in-test or variable resistor.

The parallel circuit may be set up by adjustment of R_b (Fig. 6). The required value may be calculated from Eq(2):

$$R_b = \frac{1{\cdot}25(V_{cc}-V_{be})}{c\phi},$$

or by practical adjustment of the resistor.

When setting up the circuit by practical adjustment, the resistor is set at the point where the transistor just cuts off under full illumination. Device and circuit must be under the conditions stipulated under *Calculation* (page 20); namely, maximum V_{cc}, minimum illumination, and low temperature. Variation in sensitivity may be simulated by a further 10% decrease in illumination.

SCR Inverter for Fluorescent Lamps

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Design and performance details of an SCR inverter circuit are given. The circuit operates from a 110V d.c. supply (as used in railway carriages) and feeds five 80W fluorescent lamps. The design can be adapted for other loadings.

INTRODUCTION

Silicon controlled rectifiers (SCRs) make it possible to design high-power inverters ranging from about one hundred watts to several kilowatts, using only two devices in push-pull and suitable for d.c. supply voltages of up to about 200V. High efficiency can be achieved, provided that the frequency of operation is not too high to make the turn-off time of the SCRs comparable with the half cycle time.

The SCR inverter described in this article operates at 1kc/s and is capable of providing an output power of about 400W. The inverter is specially designed for driving fluorescent lamps from 110V d.c. supplies, which are commonly used in railway carriages.

BASIC SCR INVERTER

The basic SCR push-pull inverter circuit (Ref. 1) is shown in Fig. 1. The circuit is of a driven type and will operate only if a positive trigger pulse is applied to the gate of each controlled rectifier.

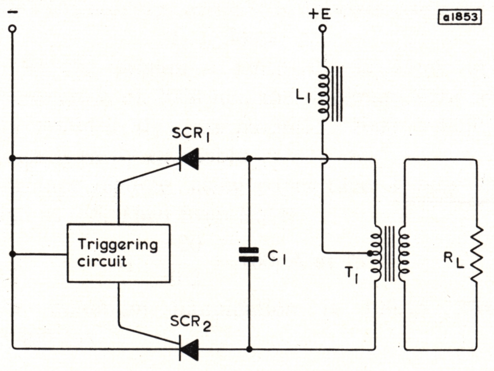

Fig. 1—Basic SCR inverter

In order to explain the operation of the inverter it is assumed that SCR$_1$ is conducting and SCR$_2$ is cut off. The anode current flowing from the supply through one half of the primary winding produces a voltage of 2E at the anode of SCR$_2$. This voltage will also appear across the commutating capacitor C$_1$.

The next input pulse will trigger SCR$_2$ into conduction, and the anode voltage of SCR$_2$ will fall to the forward conducting value. The action of the commutating capacitor causes a voltage of approximately $-2E$ to appear across SCR$_1$. The reverse voltage will be applied across SCR$_1$ long enough for it to recover to the blocking condition. The next trigger pulse will switch SCR$_1$ on again, thus completing the cycle.

The inductor L$_1$ is included to limit the current flow which charges the commutating capacitor during the switching period. The charging time has to be long enough to keep the SCR which is to be switched off, reverse biased until it recovers to its blocking state.

If the frequency of operation is low, the current and voltage waveforms are almost square waves, as switching transients occupy only a small fraction of each half cycle. Although smaller inductors and transformers may be used at higher frequencies, the switching times become a larger fraction of the total cycle time and more energy is required for commutation, resulting in worse voltage regulation and lower efficiency. There is a limit to the maximum frequency of operation. It is determined by the turn-off characteristic of the SCR and the power dissipated in the device during the switch-on transient.

The basic circuit has several disadvantages. If the load is suddenly reduced, a voltage will be induced across the inductor L$_1$, and it will also appear across the SCR which is cut off. This can be overcome by connecting a diode across L$_1$. Far worse is the problem of varying load condition, which necessitates a change in the value of the commutating capacitance for each load.

The circuit of Fig. 1 is therefore not suitable for operation with fluorescent lamps, as they present a variable load at the instant of switching on. Each lamp initially looks like an open circuit except for the very small load provided by the heaters. Also, if several lamps are used, not all lamps will strike at once; and, since it is difficult to adjust the value of the commutating capacitor instantly, there is a danger of either losing commutation voltage under heavy loads or having excessive voltage across the SCR under light loads.

IMPROVED INVERTER CIRCUIT

An improved inverter circuit (Refs. 1 and 2) is shown in Fig. 2. The inductor L$_1$ is connected between the junction of the two cathodes and the negative supply rail. Two feedback diodes D$_1$ and D$_2$ are added to limit the peak voltage across the SCRs under light-load and open-circuit conditions. The diodes prevent the voltage across each SCR from rising much above twice the supply voltage, thus allowing the use of devices with lower breakover voltage. They also compensate for reactive loads by feeding reactive power back into the supply, and they maintain a square-wave output voltage under all load conditions provided that the frequency of operation is not too high for the minimum turn-off time required. The feedback diodes are tapped across the primaries of the output transformer, otherwise a small value resistor should be included in series with each diode to limit the peak current and power dissipation, as shown in Fig. 3.

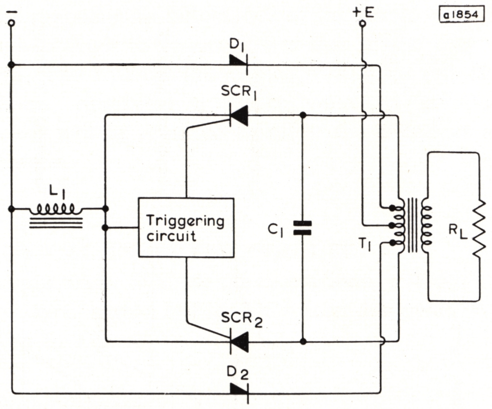

Fig. 2—Improved inverter with feedback diodes

Theory of Operation

When triggering pulses are applied in turn to the gate of each SCR in Fig. 2, the devices will be switched on and off alternately. If it is assumed that SCR$_1$ is conducting and SCR$_2$ is cut off, the anode current of SCR$_1$ flowing through one half of the primary winding will cause a voltage of 2E to appear at the anode of SCR$_2$ and across the commutating capacitor C$_1$ because of transformer action. When the next trigger pulse is applied to the gate of SCR$_2$, the SCR will be switched on and the voltage across L$_1$ will rise momentarily to twice the supply voltage. This voltage will be applied to SCR$_1$ in a reverse direction so that it will tend to cut it off; and, provided that the reverse voltage is applied for a long enough period, SCR$_1$ will recover to its blocking state. The time required is obtained by suitably choosing the values of C$_1$ and L$_1$.

With a resistive load the exchange of energy between C$_1$ and L$_1$ is oscillatory. When the anode voltage of the SCR which has been switched on falls below earth, the diode connected across the SCR will start to conduct and will damp the oscillation.

With an inductive load the load current flowing through the transformer cannot be switched off instantaneously. When SCR$_2$ is switched on and SCR$_1$ is switched off, the inductive current will continue to flow through the transformer primary and diode D$_2$ until the energy stored is transferred back into the supply. The load current then reverses. The feedback current flowing through D$_2$, however, will cause the anode current of SCR$_2$ to fall to zero, so that SCR$_2$ will have to be re-triggered when the load current reverses.

Since the triggering arrangement would become rather complex if required to provide re-triggering at the correct time for a variable load, the trigger pulses are applied for the whole of the half cycle by means of a symmetrical square-wave generator.

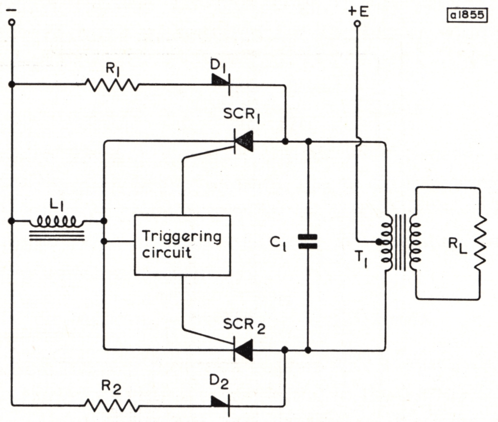

Fig. 3—Inverter with modified feedback circuits

Design Considerations

As in the case of transistor inverters (Refs. 3 and 4) the required output power, the output voltage, and the available supply voltage will be known, so that suitable SCRs may be selected.

The feedback diodes must be capable of carrying the maximum load current which they take during the commutation interval, and they must also withstand the peak inverse voltage to be encountered. Triggering requirements and suitable circuits are discussed on page 26.

An SCR can be switched on by a gate signal. It cannot normally be turned off by the gate but only by external circuit components. In inverter circuits a capacitor is used for this purpose. The capacitor must be carefully selected (Ref. 5) to have low inductance and low series resistance. Otherwise the commutation may be affected, especially at higher frequencies, or the capacitor may become overheated.

In designing SCR inverters the commutation requirements are of prime importance. The requirements must be satisfied under the worst conditions of operation, which occur under maximum load (especially when the load is inductive) and at the maximum ambient temperature.

The values of C$_1$ and L$_1$ are approximately given by

$$C_1 = \frac{t_o I_R}{E} \qquad \ldots(1)$$

and

$$L_1 = \frac{t_o E}{I_R} \qquad \ldots(2)$$

where

I_R = maximum load current referred to the primary
t_o = turn-off time
E = supply voltage.

The output transformer should be made to operate well below saturation to improve starting. If the transformer saturates, or the initial flux in the transformer happens to be near saturation when the circuit is started, the flux will be increased further and the transformer will saturate and prevent commutation when the next trigger pulse is applied.

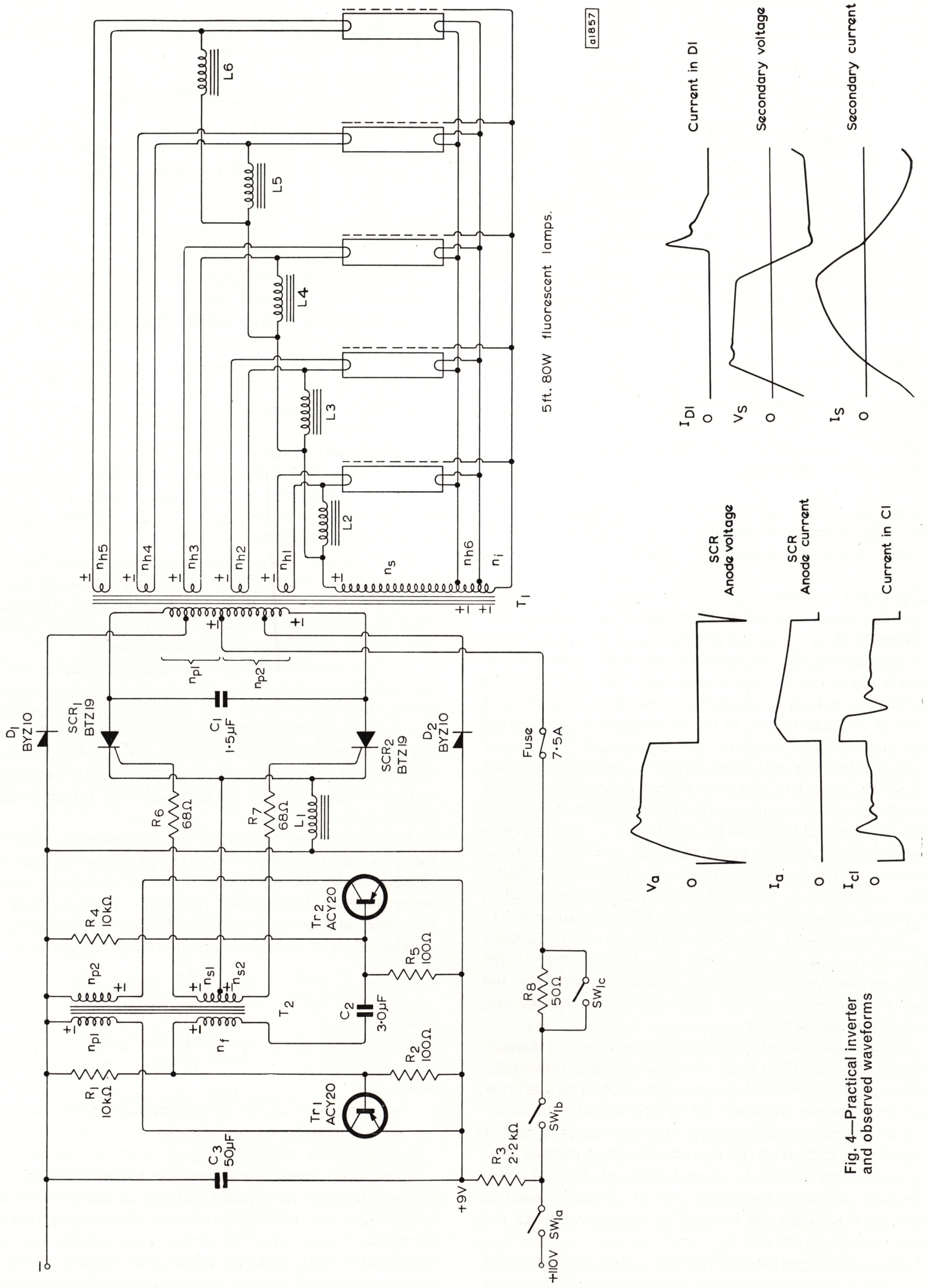

Fig. 4—Practical inverter and observed waveforms

The magnetising inductance of the transformer should be high enough to reduce losses due to the magnetising current. The leakage inductance should be reduced to a minimum, otherwise a larger value of commutating capacitor will be required to offset the effects of leakage inductance.

PRACTICAL INVERTER CIRCUIT

On the basis of these considerations, a 400W inverter of the type shown in Fig. 2 has been designed. Its circuit is given in Fig. 4.

Operating Frequency

In practice it was found that the optimum frequency was about 1kc/s. At higher frequencies the efficiency of the inverter falls off owing to the increasing ratio of the commutation time to the half cycle time and to the higher transformer loss. At lower frequencies, although the efficiency is higher, the size and values of the components (output transformer, commutation capacitor, and ballast inductors) increase excessively.

Output Transformer

The requirements of the load are considered first, in order to realise a practical inverter circuit. In the case of fluorescent lamps the load is complex, since reactive ballasting is used. If capacitive ballasts are employed, the peak current flowing through each lamp under normal working conditions is excessively high. Inductive ballasts are therefore used for all five of the 5-foot 80W lamps that have been used in a demonstration circuit. These lamps operate at 0·85A with 106V across them (Ref. 6). The starting voltage is three to four times higher, and the filament voltage is about 7V.

The output transformer T_1 (Fig. 4) is designed to allow for the above requirements of the fluorescent lamps. The primaries of the transformer are tapped, and the feedback diodes are connected as shown in Fig. 2 to reduce circuit losses by eliminating series resistors. This will result in an increased peak voltage across each SCR. The voltage will be more than twice the supply voltage by an amount depending on the positions of the tapping points. This in turn will cause the output voltage to depart slightly from a square wave. A compromise is usually necessary between the actual tapping points and the peak voltage developed across each SCR. In practice, the feedback diodes are tapped at points about 15% of the primary turns away from the anodes.

A single-loop C-core in 0·002in strip, type HWR50/32, gapped to 0·008in to prevent saturation, is used for the transformer. Each primary has 110 turns of 17s.w.g. Lewmex copper wire tapped at 17 turns. Both primaries are bifilar wound to minimise leakage inductance, and are sandwiched between the secondary, which is wound in two halves. If the voltage drop across the SCRs and the transformer primary is neglected, the number of turns required for the secondary winding (Fig. 4) is given by

$$n_s = n_p\frac{V_s}{E} = 110\frac{160}{110} = 160 \text{ turns}$$

for which 19s.w.g. Lewmex copper wire is used. The number of turns n_h required for the heater windings is given by

$$n_h = n_p\frac{V_h}{E} = 110\frac{7}{110} = 7 \text{ turns}$$

for which 23s.w.g. Lewmex copper wire is used, except for the common heater winding n_{h6} to all five lamps, for which 17s.w.g. Lewmex copper wire is used.

Finally, the number of turns n_i required for the ignition winding is given by

$$n_i = n_p\frac{V_i}{E} = 110\frac{220}{110} = 220 \text{ turns}$$

for which 32s.w.g. Lewmex copper wire is used. One end of the ignition winding is connected to the common heater winding and the other end to the earth strips of the fluorescent lamps which are all joined together, as shown in Fig. 4. This winding provides an excess voltage which is required for reliable starting of the lamps.

Fluorescent Lamp Ballasts

In order to simplify the design of the ballast inductance required for each lamp, only the fundamental component of the square-wave output voltage is considered. From a vector diagram the voltage developed across the ballast reactance V_X is given by

$$V_X = \sqrt{(V_S{}^2 - V_L{}^2)}$$

where V_S is the secondary voltage and V_L is the voltage across the lamp under normal working conditions. Therefore

$$V_X = \sqrt{(160^2 - 106^2)} = 120\text{V}.$$

For a lamp current of 0·85A the ballast reactance is

$$X_L = \frac{V_X}{I_L} = \frac{120}{0{\cdot}85} = 141\Omega,$$

hence the ballast inductance

$$L = \frac{X_L}{2\pi f} = \frac{141}{2\pi \times 1 \times 10^3} \simeq 22\text{mH}.$$

The required inductance may be obtained by using 0·002in strip C-core type HWR10/16 with a 0·008in air gap and 142 turns of 18s.w.g. Lewmex copper wire.

Commutating Components

It remains now to find the values of the commutating capacitor C_1 and the inductor L_1.

If it is assumed that the output power is 400W and the overall efficiency is 80%, the input power needed is 500W, so that for a 100V supply the load current is approximately 4·5A.

The turn-off time for the Mullard BTZ19 will be taken as 30µs for calculating the values of C_1 and L_1. From Eqs(1) and (2) and a turn-off time of 30µs, the value of the commutating capacitor C_1 is approximately 1·23µF, and the value of the inductance L_1 is 0·73mH.

Because the load is mostly inductive, owing to the ballast chokes used for the lamps, a slightly higher value of the commutating capacitance is required. It was found that with a value of 1·5µF the commutation time was approximately 30µs. The commutation time is the time available for the SCR to recover; that is, the time taken for the reverse voltage across the SCR to fall to zero.

The value of the inductor L_1 has not much effect on the commutation time. As mentioned previously, its main purpose is to provide the spike of voltage to turn off the SCR which was conducting, and to limit the peak current during the commutation period. With the above calculated value of 0·73mH the peak anode current is excessive. In order to limit it to about 10A an inductance of 4·5mH is required. This inductance may be obtained by using 0·002in strip C-core, type HWR30/16, with a 0·08in air gap to prevent the core from saturating, and 135 turns of 15s.w.g. Lewmex copper wire. This is the effective inductance required for normal working conditions with a mean current of about 4·5A, and not the self-inductance of the coil at zero d.c., which is about five times higher.

Cooling Requirements

The SCRs and the feedback diodes should be mounted on suitable heat sinks to ensure good cooling of the devices, especially at higher ambient temperatures.

The average forward power dissipation of the BTZ19 for an average anode current of 4·5A is about 7·2W, and the junction temperature rise above stud, for the device mounted without mica washer, is 3·1degC/W. With a heat sink having a thermal resistance of 3·4degC/W, the corresponding temperature rise is (3·1+3·4) 7·2 = 46·7degC. Therefore for a junction temperature of 125°C the maximum permitted ambient temperature is 125−46·7 = 78·3°C. A smaller heat sink, that is a heat sink with a higher thermal resistance, may be used if the ambient temperature is restricted to a lower figure.

Heat sinks are also necessary for the feedback diodes. Because of the lower power requirements, about 2·8W, smaller heat sinks may be used. Alternatively, the diodes may be mounted on the same heat sinks as the SCRs. In this case, mica or PVC washers must be employed, and either the maximum ambient temperature must be restricted to about 60°C, or the size of each heat sink must be increased to allow for the additional heat to be conducted away.

TRIGGERING CIRCUIT

SCR inverters are basically of the driven type; and, as mentioned on page 23, it is necessary, for proper operation, to apply triggering pulses alternately to each gate. There are both minimum and maximum values for gate excitation which must be satisfied. The minimum and maximum voltage and current values depend on the ambient temperature.

To trigger all BTZ19s within a temperature range of −55°C and 125°C, a minimum gate current of 30mA and a minimum gate voltage of 2V are required. The maximum values are 2A and 10V respectively, but the gate dissipation must be limited to 5W peak and to the maximum average power of 500mW for 20ms.

Design Considerations

The shape and duration of each pulse depend on the type of inverter circuit used, and on whether the load is purely resistive or reactive. As explained under *Theory of Operation* (page 23, a symmetrical square-wave triggering arrangement is required for the modified circuit.

Four triggering circuits were considered. These were:

(i) cross-coupled multivibrator,
(ii) emitter timed multivibrator,
(iii) push-pull blocking oscillator,
(iv) push-pull transistor inverter with capacitive timing.

The first two circuits can be made to trigger satisfactorily the inverter circuit shown in Fig. 1, for which the square-wave triggering output is differentiated and only a narrow pulse is used.

In the case of the improved inverter circuit shown in Fig. 2, the triggering pulse is applied for the whole of the half cycle. A small coupling transformer is employed to feed the trigger pulses to the gates of the SCRs. This makes the other two circuits, (iii) and (iv), more suitable, as they already include a transformer, to which windings can be added. Both circuits (iii) and (iv) provide symmetrical square-wave outputs. The push-pull transistor inverter circuit with capacitive timing, however, has several advantages over the push-pull blocking oscillator circuit. These are: the frequency of operation is less dependent on the load, the diode which is used for starting in the blocking oscillator circuit is not necessary, and the timing can be adjusted by varying either C or R.

Practical Triggering Circuit

A practical example of a triggering circuit used for the SCR inverter is given in Fig. 5. The circuit uses a linear output transformer, and the timing is controlled by a C-R combination. The complete design procedure is given in Ref. 4, and therefore it will not be repeated here. Only the necessary equations required for the design will be quoted.

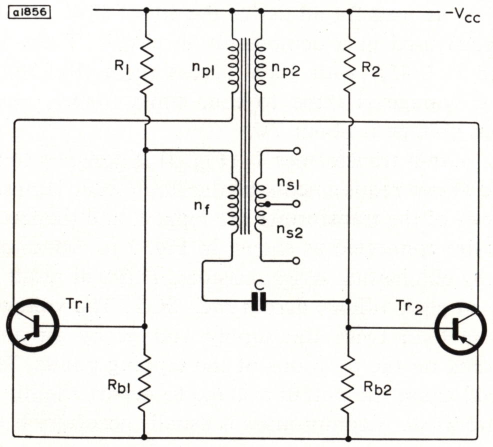

Fig. 5—Triggering circuit

It is necessary to make a few assumptions in order to design a practical triggering circuit based on the transistor inverter with C-R timing. These are: the values of the supply voltage, the magnetising current of the transformer, and the peak collector current.

The supply voltage chosen for the triggering circuit depends on the type of transistor used. For the Mullard ACY20 a supply voltage of 9V can be assumed. The magnetising current of the transformer in this case is a compromise between the value of inductance required for the primary and the peak collector current, allowing 5mA for the load current which is required for triggering. Ideally, the magnetising current should be only a small fraction of the load current. In this case, however, a rather high value of the primary inductance would be required and would result in a rather large transformer. The efficiency of the triggering circuit is of no importance, as it is working at a very low level compared with the output power of the inverter, therefore to keep the size of the triggering transformer down, a magnetising current of 200mA is assumed, which for a frequency of 1kc/s calls for a primary inductance of 22·5mH.

With two Ferroxcube E cores, type FX1105, the number of turns required for each primary is given by $19{\cdot}5\sqrt{L}$, where L is in mH, which gives 92 turns. If a 2:1 turns ratio is used between primary and secondary, 46 turns will be required for each half of the secondary. The primary and secondary use 34s.w.g. Lewmex copper wire.

The peak collector current $i_{c(pk)} = I_R + I_m$, where I_R = load current referred to the primary and I_m = magnetising current of the transformer. Therefore $i_{c(pk)} = 50+200 = 250\text{mA}$. The minimum base current required to support that collector current is given by

$$I_{b(min)} = \frac{i_{c(pk)}}{\beta} = \frac{250}{50} = 5\text{mA},$$

and the minimum current to which the timing capacitor C is discharged is

$$I_{1(min)} = I_{o(min)} + \frac{I_{b(min)}R_o + V_o}{R_b}$$

where R_o and V_o are the linearised input equivalents of the transistor. For ACY20 transistors R_o and V_o are 25Ω and 0·37V respectively. If it is assumed that $R_b = 100\Omega$

$$I_{1(min)} = 5\times10^{-3} + \frac{(5\times25\times10^{-3})+0{\cdot}37}{100} = 9{\cdot}95\text{mA}.$$

The feedback voltage required to charge the timing capacitor in order to produce the required current, assuming that the peak current discharging the capacitor $i_{1(pk)} = 5I_{1(min)}$, is given by

$$\begin{aligned} V_f &= 3I_{1(min)}R_b\left(1+\frac{R_o}{R_b+R_o}\right)+\frac{V_oR_b}{R_b+R_o} \\ &= 3\times9{\cdot}95\times10^{-3}\times100\left(1+\frac{25}{100+25}\right)+ \\ &\qquad +\frac{0{\cdot}37\times25}{100+25} \simeq 3{\cdot}66\text{V}. \end{aligned}$$

Thus the number of turns required for the feedback winding, which is given by $n_f = n_p(V_f/V_{cc})$, is approximately 38 turns, for which 30 s.w.g. Lewmex copper wire is used.

SWITCHING PRECAUTIONS

If the d.c. supply is obtained from a.c. mains and a rectifier circuit, then certain precautions must be taken when the inverter is switched on or off.

Switching off the a.c. supply will not give an instantaneous reduction of the d.c. voltage from 110V to zero, because of the time-constant of the smoothing circuit. When the d.c. voltage has decayed to one-third of its value, the supply to Tr_1 and Tr_2 will be 3V, which is insufficient to allow operation of the triggering circuit. There will be no triggering pulses, and the reverse voltage necessary to switch off the conducting SCR will not appear.

The supply to the inverter, however, will be about 38V, which is sufficient to maintain in conduction the SCR that was last 'on'. This SCR (for example, SCR_1 in Fig. 4) will remain in conduction, in the absence of the negative switch-off voltage, until the d.c. voltage has fallen almost to zero volts.

If, during this period, the a.c. supply is switched on again, there is a danger that SCR_2 will be triggered before there is sufficient negative voltage to switch off SCR_1. The SCRs will then be conducting together. Commutation will be lost, and there is a possibility of damage to one or both of the SCRs.

To avoid this situation it is essential that, where a rectified a.c. supply is used, the d.c. line to the SCRs shall be broken before switching off the a.c. input to the rectifier unit; and a delay of a few seconds must be allowed before the a.c. is switched on again, in order that the smoothing capacitors shall have time to discharge completely. Switch SW_1 is to be opened before the a.c. supply is switched off, and it is to be closed after the a.c. is switched on. The three contacts are arranged to close in the order a, b, c, and to open in the order c, b, a.

Protection against incorrect switching is provided by the 7·5A fuse in the d.c. line. It also provides protection against short-circuits and against failure of the triggering circuit.

PERFORMANCE OF COMPLETE INVERTER

The complete circuit of the 400W inverter, with its triggering arrangement, is shown in Fig. 4 (page 24). The supply voltage for the triggering circuit is derived from the 110V line by means of a decoupling resistor R_3 and capacitor C_3. Resistors R_6 and R_7 are included in each gate lead to limit the maximum gate current.

Although the inverter operates satisfactorily under no-load conditions, its power consumption is high. Therefore it should not be used without load for long periods, as the power drawn from the supply is all dissipated in the circuit, with a high proportion of it in the inductor L_1—which may get rather hot.

The performance of the inverter is summarised in Table 1.

TABLE 1
Performance of Inverter

	Open-circuited load	Full load	
Supply voltage	110	110	(V)
Supply current	1·32	4·35	(A)
Input power	145	478	(W)
Commutation time	53	30	(μs)
Frequency	—	1	(kc/s)
Output power	—	400	(W)
Relative efficiency	—	100	(%)

The relative efficiency is the ratio of (a) the input power required for the lamps from 50c/s mains at 230V to (b) the input power required for the inverter (including the triggering circuit) from a 110V d.c. supply, the light output from the lamps being the same in each case.

Symmetrical triggering waveforms are applied to the gates of the SCRs. The SCR anode voltage and current waveforms on full load are shown in Fig. 4. The current waveforms in the commutating capacitor C_1 and the feedback diode D_1, and the waveforms of the secondary voltage and current of T_1, are also included.

As the inverter is of a driven type, the frequency of operation is not affected by the load conditions. The shape of the waveforms, however, and the commutation time of the circuit, vary with the load to some extent. The commutation time may be increased, if required, by increasing the value of the commutating capacitor. This, however, increases power consumption under no-load conditions and reduces the efficiency of the inverter on full load. The efficiency of the inverter also increases with the load, and is highest at maximum load.

Although 5ft 80W fluorescent lamps are used for the demonstration circuit, other types of fluorescent lamp can be used, provided that the necessary adjustments are made to the secondary of the transformer and to the ballasting components for the lamps.

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30Mc/s Wideband Transistor Amplifier

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The design of a 30Mc/s wideband transistor amplifier is described. The gain is 60dB and bandwidth 15Mc/s. Three different types of circuit are discussed, together with the simple theory of single-tuned coupling circuits. The circuit of a stagger-tuned amplifier is given with practical design details. The effects of temperature and transistor parameter spreads are discussed.

INTRODUCTION

The increasing use of shorter pulses in radar systems involves the design of video and i.f. circuits with a wide bandwidth. For a particular application a broadband i.f. amplifier was required with a gain of 60dB and a bandwidth of 15Mc/s centred on 30Mc/s.

Four methods of designing such an amplifier are considered:

(i) A broadband video amplifier preceded or followed by a bandpass filter.
(ii) A synchronously-tuned amplifier with one tuned circuit per stage. Each stage is tuned to the same frequency.
(iii) A double-tuned amplifier with two tuned circuits per stage. The stages are tuned to the same frequency.
(iv) A stagger-tuned amplifier with successive stages tuned to different frequencies. Individual stages may be single- or double-tuned.

It will be shown in the next section that a stagger-tuned amplifier is suitable for the present application. The use of single-tuned stages in such an amplifier enables the circuit design to be simpler and less critical than that of the double-tuned amplifier. The theory of single-tuned broadband interstage coupling is discussed and compared with results obtained in practice. The circuit diagram of a practical amplifier is shown in Fig. 8 (page 32).

FOUR TYPES OF WIDEBAND AMPLIFIER

The power gain of an amplifier, expressed in dB, is defined as

$$10\log_{10}\frac{\text{power delivered to load}}{\text{power absorbed by amplifier input circuit}}.$$

Voltage gains are considered throughout this article, and since these have been measured between points of equal impedance it is permissible to express the gain in dB as

$$20\log_{10}\frac{\text{output voltage}}{\text{input voltage}}.$$

The bandwidth is defined as that frequency range over which the power gain remains within 3dB of its value at band centre. The edges of the band are therefore the points at which the voltage gain has fallen to 0·707 of the value at band centre.

Broadband Video Amplifier and Bandpass Filter

The bandwidth of the video amplifier is determined by the upper frequency limit of the required passband. In an amplifier with 15Mc/s bandwidth centred on 30Mc/s the gain will be 3dB below its maximum value at 37·5Mc/s. Hence the minimum bandwidth of the video amplifier is 37·5Mc/s, which is considerably greater than the 15Mc/s bandwidth of a tuned amplifier. Since the gain-bandwidth product of a stage is fixed, the video stages would have less than half the gain of the tuned stages, so that this system is wasteful. A video amplifier and bandpass filter would be suitable for a broadband amplifier at a lower centre frequency.

Synchronously-tuned Amplifier

With synchronously-tuned stages there is a maximum bandwidth which can be achieved for any given gain. This occurs when the individual stage gain is $\sqrt{e}$(1·65 or 4·3dB). Hence, for a gain of 60dB the maximum bandwidth is obtained when the number of stages is $60/4{\cdot}3 = 14$. If a gain-bandwidth product of 100Mc/s is assumed for a transistor, the bandwidth of a single stage would be $100/1{\cdot}65 = 60$Mc/s. For n stages in cascade this bandwidth is reduced by a factor of $1/(1{\cdot}2\sqrt{n})$. Hence, when $n = 14$ the overall bandwidth is $60/(1{\cdot}2\sqrt{14})$ or 13·3Mc/s. Thus it is not possible to obtain 60dB gain and 15Mc/s bandwidth with transistors whose gain-bandwidth product is 100Mc/s.

Double-tuned Amplifier

The gain-bandwidth product of a single-tuned stage may be improved by the use of double-tuned circuits. Both the primary and secondary windings are tuned to the centre frequency, and the coupling is designed to give a maximally flat response.

The inductance of the secondary winding depends on the primary inductance and turns ratio, which are determined by the gain, bandwidth, and centre frequency of the stage. If this inductance is small, the secondary winding may be tuned directly with the base input capacitance of the following stage and an additional capacitance (Fig. 1). If the secondary inductance is large, the base must be connected to a tap on the secondary winding which is tuned with an additional capacitance (Fig. 2).

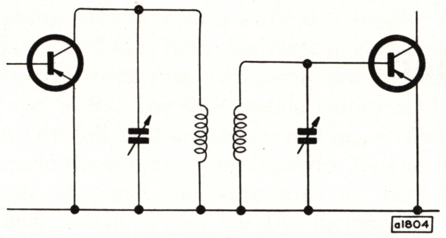


Fig. 1—Double-tuned circuit with small secondary inductance

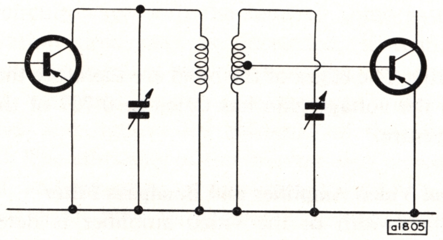


Fig. 2—Double-tuned circuit with large secondary inductance

The first method of tuning (Fig. 1) was practicable above 40Mc/s. Below this frequency, the tapped secondary (Fig. 2) was more suitable. Hence below 40Mc/s the design of double-tuned circuits is considerably more complicated than that of the corresponding single-tuned circuits. This is also seen in the theoretical calculations below. However, a double-tuned amplifier may be necessary if an accurate control of the passband is required.

Stagger-tuned Amplifier

The stagger-tuned amplifier, like the double-tuned amplifier, will give a greater gain-bandwidth product than the synchronously-tuned amplifier. The stagger-tuned amplifier has been chosen for the present application because it is possible by this means to obtain the required gain and bandwidth by means of single-tuned circuits.

In order to simplify the tuning a stagger-tuned triple was used. One stage is tuned to the centre frequency, a second stage above, and a third stage below. More than three stages were required to obtain sufficient gain, and the present amplifier consists of one stage tuned to the centre frequency, two stages above, and two stages below.

The design of stagger-tuned stages is described by Valley and Wallman (Ref. 1). If the bandwidth of the present amplifier is 15Mc/s, one stage will be tuned to 30Mc/s with 15Mc/s bandwidth, two stages tuned to 36·8Mc/s with 8·5Mc/s overall bandwidth, and two stages tuned to 25·4Mc/s with 5·6Mc/s overall bandwidth. The bandwidth contraction for two stages is 0·64. Hence, each higher frequency stage will have a bandwidth of 8·5/0·64 = 13·3Mc/s, and each lower frequency stage 5·6/0·64 = 8·7Mc/s. The order of the stages is shown in Fig. 3.

DESIGN OF BROADBAND SINGLE-TUNED INTERSTAGE COUPLING NETWORKS

This section describes the design of single-tuned interstage coupling networks which may be used in synchronous or stagger-tuned amplifiers. The interstage coupling network must determine the centre frequency and bandwidth of the stage, while transferring sufficient r.f. power from the collector to the following base to maintain the stage gain. The output impedance of one transistor must be matched to the input impedance of the next, the former being about ten times the latter. Matching by inductive coupling is more efficient than capacitive coupling and was therefore adopted. A typical single-tuned coupling network is shown in Fig. 4, and will be considered in more detail.

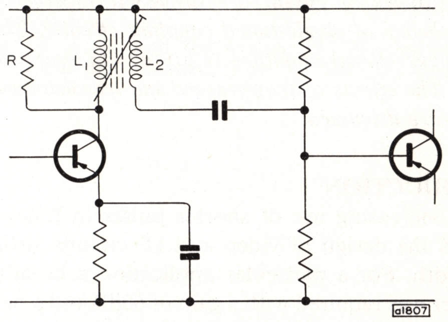


Fig. 4—Single-tuned interstage coupling network

The coil L_1 is tuned to the centre frequency by the reflected capacitance from the following base. Tuning is effected with a screw dust core, and an additional variable capacitor is not necessary. The resistance R damps the circuit to the required bandwidth.

Theoretical Considerations

The circuit shown in Fig. 4 may be replaced by the equivalent circuit shown in Fig. 5. In this circuit

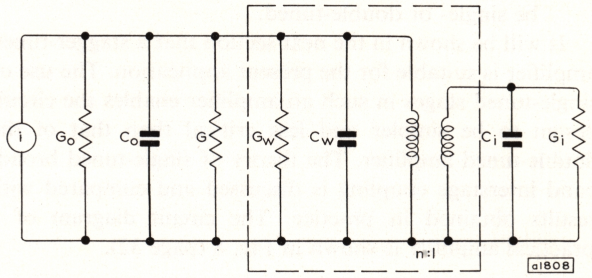


Fig. 5—Equivalent circuit of a single-tuned interstage coupling network

G_o, C_o are the output conductance and capacitance of the first transistor.

G_i, C_i are the input conductance and capacitance of the next stage.

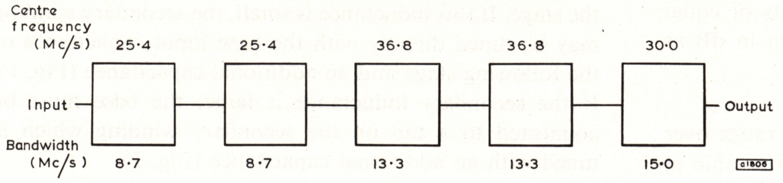


Fig. 3—Stages of stagger-tuned amplifier

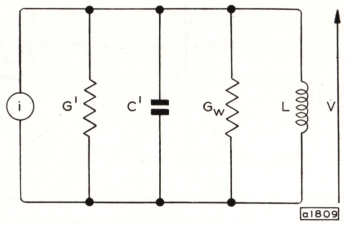
G_w, C_w are the effective loss conductance and wiring capacitance of the coil.

i is an r.f. current generator replacing the first transistor.

ω is the centre frequency of operation.

It is proposed to determine the voltage across the primary coil and hence the voltage at the input of the following stage. The coupling coefficient of the coil is assumed to be high, and feedback through the transistor is not considered.

Fig. 6—Simplified primary circuit



The primary circuit may be simplified as shown in Fig. 6. In this circuit

G′ is the total conductance across the coil (excluding coil loss).

C′ is the total capacitance.

The voltage V generated across the coil is given by

$$V = \frac{i}{G'+G_w+j\left(\omega C' - \dfrac{1}{\omega L}\right)}.$$

At the centre frequency, $C' = 1/\omega L$.
Therefore

$$V = \frac{i}{G'+G_w}. \qquad \ldots(1)$$

The bandwidth of the circuit is given by

$$B = \frac{G'+G_w}{2\pi C'}. \qquad \ldots(2)$$

In narrow-band circuits B is small and additional damping is kept to a minimum. Damping from coil losses predominates, so that $G_w \gg G'$ and

$$V = \frac{i}{G_w}. \qquad \ldots(3)$$

In broadband circuits additional damping must be provided to increase B, and this may swamp the effect of the coil. In this case,

$$G' \gg G_w \quad \text{and} \quad V = \frac{i}{G'}. \qquad \ldots(4)$$

The Q factor of the coil is given by $1/(\omega L G_w)$, hence $G_w = L/(\omega LQ)$ and Eq(1) becomes

$$V = \frac{i}{G' + \dfrac{1}{\omega LQ}}. \qquad \ldots(5)$$

Hence, in narrow-band circuits

$$V = i\omega LQ \qquad \ldots(6)$$

and in broadband circuits, as already given (Eq. 4)

$$V = \frac{i}{G'}.$$

It will be shown later that the effective damping is about 500Ω for typical stages, therefore $G' = 1/500$mho. For the coils in use, $Q = 120$ and $L = 2\mu H$. The centre frequency is 30Mc/s, therefore

$$\frac{1}{\omega LQ} = \frac{1}{4{\cdot}3\times10^4}\text{mho.}$$

Hence it is seen that for typical broadband circuits $G' \gg 1/(\omega LQ)$ and the external damping predominates. Now

$$i = g_m V_{in}$$

where

g_m = mutual conductance of the first transistor
V_{in} = the input voltage on its base.

The voltage on the base of the following transistor is V/n. Hence the stage gain is $V/(nV_{in})$ or Vg_m/ni.

From Eq(4)

$$\frac{V}{i} = \frac{1}{G'}$$

and

$$\text{stage gain} = \frac{g_m}{G'n}. \qquad \ldots(7)$$

From the original equivalent circuit

$$G' = G+G_o+\frac{G_i}{n^2} \qquad \ldots(8)$$

$$C' = C_o+C_w+\frac{C_i}{n^2}. \qquad \ldots(9)$$

Substituting these into (7) and (2) gives two relations. Stage gain is given by

$$A = \frac{g_m}{n(G+G_o)+G_i/n}. \qquad \ldots(10)$$

Stage bandwidth, neglecting G_W, is given by

$$B = \frac{G+G_o+G_i/n^2}{2\pi(C_o+C_w+C_i/n^2)}. \qquad \ldots(11)$$

In practice, the maximum gain is required for a given bandwidth with G and n variables. It is desired to know the values of G and n which give a maximum value of A, where

$$A = A(G,n)$$

and a constant

$$B = B(G,n).$$

These equations apply to coils with coupling coefficients approaching unity, and must be modified for coils with realistic coupling coefficients of about 0·5. The stage gain increases with coupling coefficient, and so for single-tuned circuits the maximum attainable coupling coefficient is desirable. Hence Eqs(10) and (11) could be modified to include the effects of a constant coupling coefficient, and the problem is still the same as that formulated above. The solution of this problem is beyond the scope of this article.

A similar problem arises in the design of broadband double-tuned circuits. The design of these circuits for a given bandwidth has been established, but it has not been shown that this design gives maximum power transfer between two unequal impedances, which is a condition for maximum gain in transistor amplifiers. This problem might be approached in a similar manner to that outlined for single-tuned circuits, although there is considerably more complication, since the shape of the passband must be considered.

Practical Design

Broadband single-tuned coupling circuits, suitable for synchronous and stagger-tuned amplifiers, have been investigated experimentally by observing the gain and bandwidth of a single stage as the damping and primary/secondary turns ratio were varied. The circuit shown in Fig. 4 has separate primary and secondary windings and was simplified by feeding the base from a tap on the primary winding (Fig. 7). The primary/secondary turns ratio n was easily altered by shifting the position of the tap.

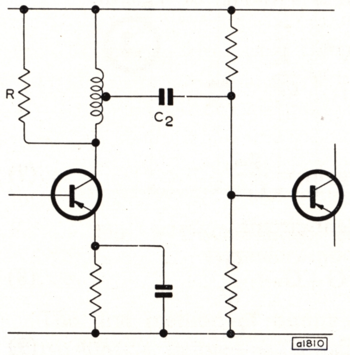


Fig. 7—Practical single-tuned interstage coupling network

It was observed that a greater gain could be achieved for a given bandwidth by the use of the circuit shown in Fig. 7 than by direct coupling between base and collector with a capacitor of optimum value (n = 1). The optimum position of the tap was not critical, and a value of n = 2 was chosen as suitable for all stages.

When n was fixed it was desirable to vary the capacitance across the secondary circuit by adjusting the series capacitor between the tap and the base (C_2 in Fig. 7). As C_2 is reduced from a high value, the gain will drop and the secondary capacitance and conductance will be reduced. The magnitudes of the capacitance and conductance and the reduction in gain are calculated in the Appendix (page 34).

For the 25Mc/s coil in the present amplifier, C_2 = 100pF.

The base input capacitance is 40pF, and the input resistance is 80Ω.

The capacitance appearing between the tap on the primary coil and earth is 17pF, and the resistance is 240Ω.

The transistor output resistance is about 4kΩ, and external damping is 1·5kΩ, giving about 1kΩ primary damping resistance. The primary and secondary resistances are therefore approximately in the ratio of n^2:1, which is the condition for maximum power transfer. The total primary damping resistance, including the reflected resistance, is about 500Ω.

This type of coupling network is used between the first four stages shown in the circuit diagram, Fig. 8.

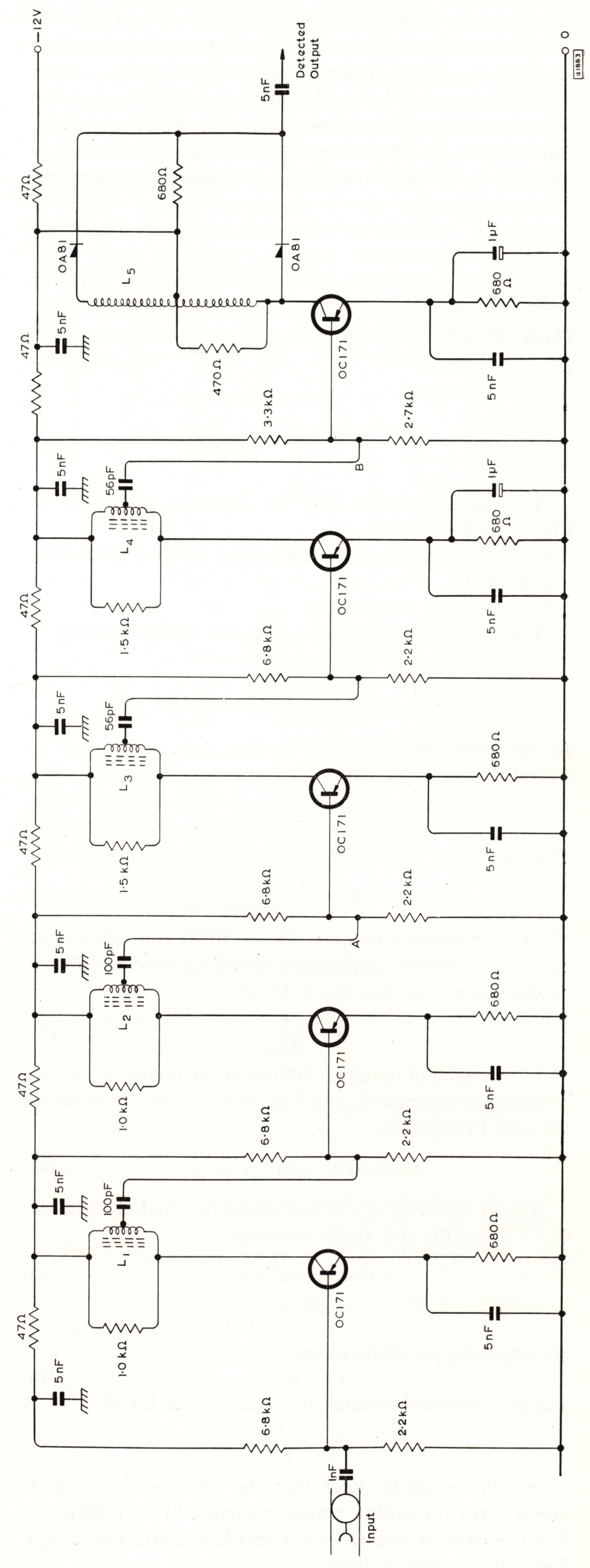


Fig. 8—Circuit diagram of the complete amplifier with a detector at the output
L1, L2: wound with 12 turns of 30 s.w.g. enamelled copper wire on Aladdin ¼in. former, centre-tapped, ⅝in. dust core.
L3, L4: as L1, L2, but 9 turns.
L5: as L1, L2, but 20 turns.

Output Stage of the Stagger-tuned Amplifier

The output stage is more heavily damped, so that the frequency response does not depend significantly on the addition of small capacitances (< 20pF) to the collector circuit.

Fig. 8 shows the complete amplifier with a balanced detector on the output. The balanced circuit helps to reduce the r.f. content in the video output. Fig. 9 shows the last stage delivering an r.f. output with 75Ω impedance. Some additional tuning is then necessary on the secondary coil, as shown. This output may be connected to another transistor stage without the additional tuning capacitor.

The modification shown in Fig. 10 enables the circuit to be used as a limiting amplifier in an f.m. system.

Fig. 9—Output stage modified to obtain a low impedance r.f. output.
L5: wound with 10 turns of 30 s.w.g. enamelled copper wire on Aladdin ¼in. former, centre-tapped

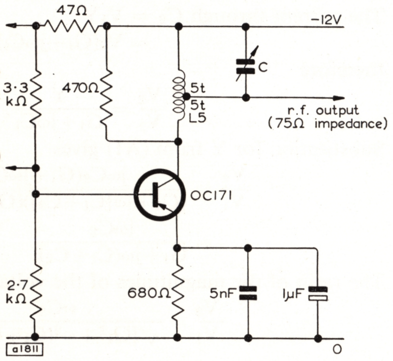
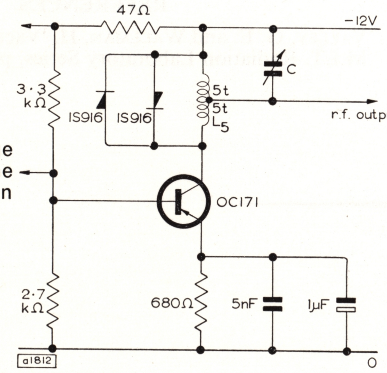


Fig. 10—Output stage modified to enable the amplifier to limit in an f.m. system



ALIGNMENT

If a 30Mc/s stage is connected to a signal generator, and a voltmeter is placed at the output, a voltage maximum is obtained by adjustment of the dust core. However, owing to the decrease of transistor gain with frequency, the maximum gain will occur slightly below the signal generator frequency. Therefore, if a maximum gain is required at any particular frequency, the dust core must be tuned for a maximum at a slightly higher frequency (for example, about 2Mc/s higher at 30Mc/s). The procedure given below allows for this, and so the dust core should be tuned for a maximum at the frequency given.

In tuning a broadband stage it is not apparent exactly where the voltage maximum occurs. The method adopted here is to screw in the dust core completely, and to screw it out slowly until the output voltage perceptibly falls. This ensures consistent results in tuning.

Tuning Procedure

Connect the signal generator to the input, and the r.f. voltmeter to point A (Fig. 8).

Tune the first two stages to 25Mc/s.

Connect the voltmeter to point B.

Tune the next two stages to 41Mc/s.

Screw the dust core on the last stage completely in.

EFFECTS OF TRANSISTOR PARAMETER SPREADS

Spreads of about 30% in the input and output admittance parameters are expected. Spreads in the input and output capacitances of the transistor are taken up in the initial tuning. If a transistor is changed, the maximum change in stagger frequency is 2Mc/s, which could alter the overall bandwidth by 1Mc/s; hence it is probably not necessary to re-align after changing transistors.

The response is not critically dependent on stage damping, and so spreads in transistor input and output conductance have a negligible effect. A spread of about ±0·75dB in stage gain is expected, giving about ±4dB for the complete amplifier.

PERFORMANCE

The following measurements were made on three similar amplifiers which had been aligned in the manner described:

Gain	Bandwidth
64dB	14·8Mc/s
71dB	15·6Mc/s
65dB	15·4Mc/s

The transfer characteristic of the amplifiers is shown in Fig. 11, and the frequency response in Fig. 12.

With a rise in the ambient temperature from 20°C to 50°C the amplifiers showed a 2% reduction in gain (dB) and a 1Mc/s reduction in bandwidth. The noise figure was 4·5dB.

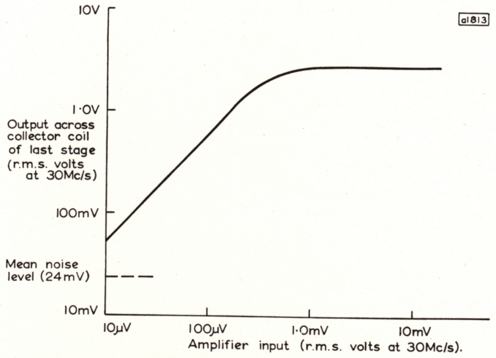


Fig. 11—Transfer characteristic of the amplifiers

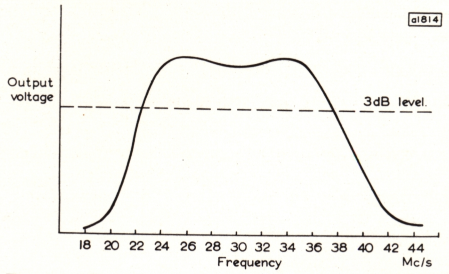


Fig. 12—Frequency response of the amplifiers

APPENDIX

The Input Admittance of a Transistor with Capacitive Coupling

The equivalent input circuit of a transistor is shown in Fig. 13 with a coupling capacitor C_2 (as shown on page 32). The capacitance and conductance of this circuit are calculated together with the fractional drop in voltage across the coupling capacitor. V_1 is the input voltage and V_2 is the voltage on the base of the transistor.

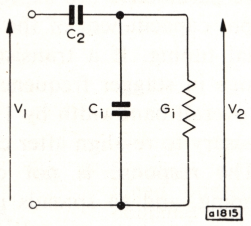


Fig. 13—Equivalent circuit of transistor with coupling capacitor

The input impedance of the circuit is

$$\frac{1}{j\omega C_2}+\frac{1}{G_i+j\omega C_1}$$

or

$$\frac{G_i+j\omega(C_i+C_2)}{j\omega C_2(G_i+j\omega C_i)}.$$

Hence the input admittance, Y, is

$$\frac{j\omega C_2(G_i+j\omega C_i)}{G_i+j\omega(C_i+C_2)} \qquad \ldots(\text{A1})$$

or

$$\frac{(-\omega^2C_iC_2+j\omega C_2G_i)(G_i-j\omega[C_i+C_2])}{(G_i+j\omega[C_i+C_2])(G_i-j\omega[C_i+C_2])}$$

or

$$\frac{-G_i\omega^2C_iC_2+\omega^2C_2G_i(C_i+C_2)+j\omega[C_2G_i^2+\omega^2C_iC_2(C_i+C_2)]}{G_i^2+\omega^2(C_i+C_2)^2}$$

or

$$\frac{\omega^2C_2^2G_i+j\omega C_2[G_i^2+\omega^2C_i(C_i+C_2)]}{G_i^2+\omega^2(C_i+C_2)^2}.$$

Therefore the input conductance is

$$\frac{\omega^2C_2^2G_i}{G_i^2+\omega^2(C_i+C_2)^2}$$

or

$$\frac{G_i}{\left(\frac{G_i}{\omega C_2}\right)^2+\left(1+\frac{C_i}{C_2}\right)^2}. \qquad \ldots(\text{A2})$$

The input capacitance is

$$\frac{C_2[G_i^2+\omega^2C_i(C_i+C_2)]}{G_i^2+\omega^2(C_i+C_2)^2}. \qquad \ldots(\text{A3})$$

The current through $C_2 = V_1Y$

$$= V_2(G_i+j\omega C_i)$$

therefore

$$\frac{V_2}{V_1}=\frac{Y}{G_i+j\omega C_i}.$$

Substitution for Y from (A1) gives

$$\frac{V_2}{V_1}=\frac{j\omega C_2(G_i+j\omega C_i)}{(G_i+j\omega[C_i+C_2])(G_i+j\omega C_i)}$$

$$=\frac{j\omega C_2}{G_i+j\omega(C_i+C_2)}.$$

The ratio of the magnitudes of the voltages is then

$$\frac{V_2}{V_1}=\frac{\omega C_2}{\sqrt{[G_i^2+\omega^2(C_i+C_2)^2]}}. \qquad \ldots(\text{A4})$$

REFERENCES

1. VALLEY, G. E. and WALLMAN, H. 'Vacuum Tube Amplifiers', M.I.T. Radiation Laboratory Series, p. 176.